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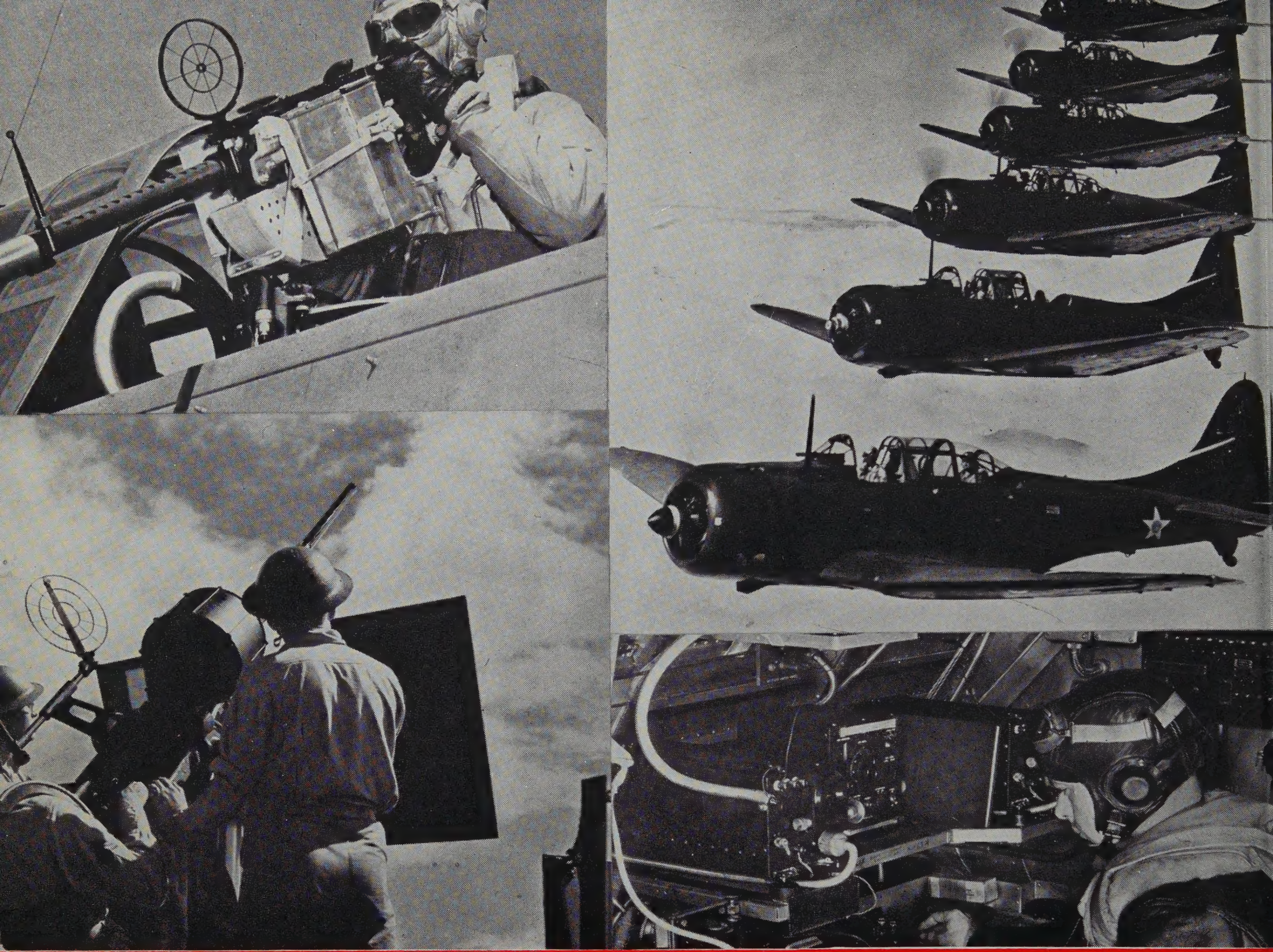

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An Electronic Potentiometer*

M. A. HONNELL,† ASSOCIATE, I.R.E.

Summary—*A direct-current, degenerative, slide-back, vacuum-tube voltmeter employing standard potentiometer design principles is described. The voltmeter is completely self-calibrated upon construction. By adjusting the voltmeter to the same reference balance condition for all voltage measurements, the grid current is readily reduced to less than* 10^{-9} *ampere. Voltages in the range from 1 to 100 volts are read to four significant figures in the experimental model of the voltmeter.*

Introduction

THE most accurate voltage standard available to the independent investigator is a standard cell employed with an accurate potentiometer. That is, a known voltage and standard resistances are utilized to establish definitely known voltages. This idea is incorporated in a direct-current vacuum-tube voltmeter which employs an accurate potentiometer and a known current, or a known voltage, as an integral part. It is a degenerative, slide-back, triode voltmeter designed to utilize the advantages of both the direct-current slide-back[1] and degenerative voltmeters.[2] A direct-current vacuum-tube voltmeter of this type is the nucleus of an accurate all-purpose instrument which may be used as a direct-current ammeter when provided with shunts, or as a radio-frequency voltmeter[2] and ammeter[3] when used in conjunction with a diode voltmeter.

Theory of Operation

Fig. 1 is the fundamental functional circuit of the voltmeter. Initially, the push-button switch PB is pressed to the position marked "set," and the cathode resistor R_k is adjusted so that the milliammeter reads exactly 0.001 ampere. Assume, then, that the voltage to be measured E_i is precisely 40 volts, and that this voltage is applied to the voltmeter with positive polarity on the grid of the tube. When the push button is released to the position marked "read," the cathode current, as indicated by the milliammeter, will change. It is apparent that the meter reading may be returned to 0.001 ampere at one particular setting of the 100,000-ohm potentiometer marked P. At this setting, R_1 is precisely 40,000 ohms, and the voltage drop across R_1 is 40 volts.

With the circuit constants shown in Fig. 1, and a reference current of 0.001 ampere, direct voltages up to 100 volts may be measured. For example, if the unknown voltage is 95 volts, the meter reading will be returned to 0.001 ampere when R_1 is 95,000 ohms. The voltage drop across R_1 is then 95 volts. For any particular adjustment of the potentiometer, $R_1+R_2=100{,}000$ ohms. It is evident that if the meter is accurately calibrated at the single point of 0.001 ampere, the unknown voltage is read in terms of the resistance of the section of the potentiometer marked R_1 in Fig. 1. Furthermore, the circuit is adjusted to identical values of element potentials and currents for all voltage measurements.

With a plate-supply voltage of 250 volts, the total voltage drop across the potentiometer is 100 volts, when the cathode current is adjusted to the reference value of 0.001 ampere. The remaining voltage drop is

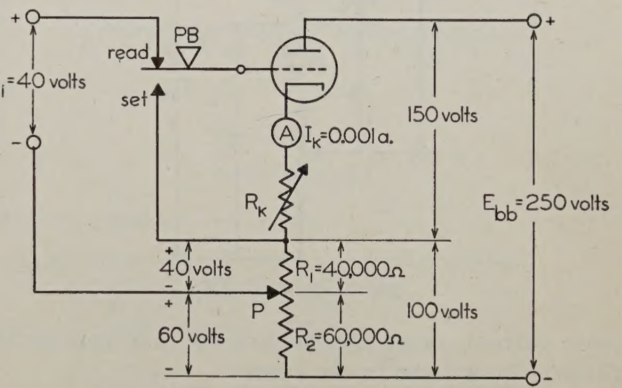


Fig. 1—Fundamental functional diagram.

across the tube and R_k. The value of the plate-supply voltage is dictated by the range of voltages to be measured and by the circuit constants employed. However, it need be only high enough to permit adjustment of the cathode current to 0.001 ampere.

One of the principal factors governing the accuracy of the voltmeter is the precision with which the cathode current may be adjusted to the reference value of 0.001 ampere. It is evident that the total resistance included in the cathode circuit has the predominating effect on the value of the plate, or cathode, current. The effect of the various circuit parameters on the sensitivity of adjustment of the plate current to the nominal value may be determined to a first approximation through a mathematical investigation of the circuit by use of the equivalent plate-circuit theorem. Referring to the basic circuit of Fig. 2, the voltage increment ΔE_g, acting from the grid to the cathode, causes an incremental change in plate current

$$\Delta I_p = \frac{\mu \Delta E_g}{r_p + R_t} \tag{1}$$

where

μ = the amplification factor of the tube,
r_p = the dynamic plate resistance of the tube, and
R_t = the total resistance in the external plate-to-cathode circuit.

The change in plate current ΔI_p flowing through R_f produces the voltage drop $\Delta I_p R_f$ with a polarity

* Decimal classification: R243.1. Original manuscript received by the Institute, September 9, 1940; revised manuscript received, June 17, 1942.

† Georgia School of Technology, Atlanta, Georgia.

[1] H. J. Reich, "Theory and Application of Electron Tubes," McGraw-Hill Book Company, New York, N. Y., 1939, pp. 571–574.

[2] W. N. Tuttle, "Type 726-A vacuum-tube voltmeter," *Gen. Rad. Exp.*, vol. 11, pp. 1–6; May, 1937.

[3] D. B. Sinclair, "The type 726-A vacuum-tube voltmeter as a radio-frequency ammeter," *Gen. Rad. Exp.*, vol. 13, pp. 1–4, August–September, 1938.

opposite to ΔE_i, the incremental change in input voltage. The net incremental grid voltage acting from the grid to the cathode of the tube is defined by

$$\Delta E_g = \Delta E_i - \Delta I_p R_f. \quad (2)$$

Substitute (2) in (1) and solve for ΔI_p to obtain

$$\Delta I_p = \frac{\Delta E_i}{\dfrac{r_p + R_t}{\mu} + R_f}. \quad (3)$$

Since both μ and r_p may be considered constant when the point of operation is in the vicinity of the nominal plate current, or as the balance point is approached, (3) may be written in the form

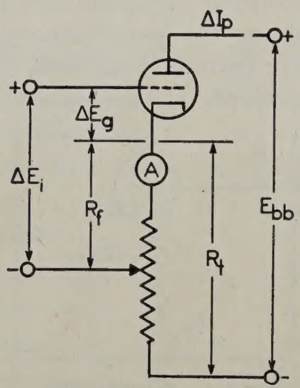


Fig. 2—Basic circuit.

$$\Delta I_p = \frac{\Delta E_i}{k + R_f} \quad (4)$$

where

$$k = \frac{r_p + R_t}{\mu}.$$

Equation (4) gives the incremental change in plate current about the balance point to a good degree of accuracy. For a given change in input voltage, the change in plate current approaches a maximum as the resistance of R_f approaches zero. Therefore, as R_f is increased, the least perceptible change in plate current is decreased for a given change in input voltage. However, the resistance of R_f is small when low voltages are measured, and it is large when relatively high voltages are measured. The effect of this is to render the per cent accuracy of adjustment of the voltmeter reasonably constant over most of its range.

Wide changes in plate voltage have no effect on the accuracy of the voltmeter so long as the plate-supply voltage remains constant during the interval between the initial and the final adjustments of the potentiometer. The degeneration present with any particular adjustment of the potentiometer aids in providing good stability of setting.

Vacuum tubes with far-different characteristics may be interchanged in the voltmeter without affecting the calibration, since the calibration of the voltmeter is dependent solely upon the calibration of the potentiometer and of a single point on the milliammeter. It is to be noted that this voltmeter differs from the usual self-calibrating slide-back vacuum-tube voltmeter in that it employs a single meter, or indicating device.

The Practical Voltmeter

In order to reduce to a minimum the manual operations necessary in making a voltage measurement, the basic circuit of the vacuum-tube voltmeter is rearranged as in Fig. 3. The push button *PB* is pressed, and $R_k - R_k'$ is adjusted so that the meter in the cathode lead reads exactly 1 milliampere. Then the push-button is released and the voltmeter is ready for use. The cathode resistor $R_k - R_k'$ consists of a fine and a coarse adjustment in order that the initial cathode-current adjustment may be precisely made.

A vacuum-tube voltmeter utilizing the circuit of Fig. 3 with a range of 0 to 100 volts is illustrated in Fig. 4. The tube employed in the voltmeter is a 6F5GT, a high-μ triode with a small glass envelope. The grid lead of the 6F5GT is brought through the top of the glass envelope, thus insuring a high-resistance path between the grid and the other elements in the tube.

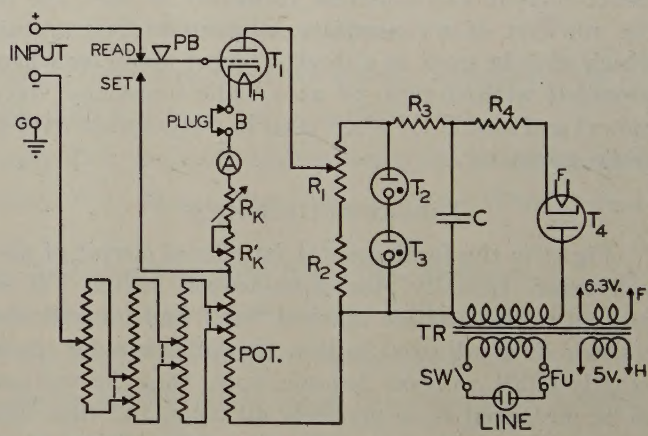


Fig. 3—The practical vacuum-tube voltmeter circuit.
R_1 = 5000-ohm potentiometer
R_2 = 25,000-ohm bleeder resistor
R_3 = 5000-ohm resistor
R_4 = 5000-ohm adjustable resistor
R_k = 250-ohm variable resistor
R_k' = resistor adjustable in ten 200-ohm steps
POT. = 100,000-ohm Thomson-Varley potentiometer
C = 16-microfarad electrolytic condenser
TR = small power transformer
T_1 = 6F5GT triode
T_2 and T_3 = VR150-30 glow-discharge tubes
T_4 = 1-V rectifier

In order that the setting of the potentiometer, and hence the unknown voltage, may be read to four significant figures, a 4-dial, 100,000-ohm potentiometer, or calibrated voltage divider, of the Thomson-Varley[4] type composed of resistors of 0.1 per cent accuracy is employed. The total resistance of this type of potentiometer remains constant irrespective of the position of the individual dials. Fig. 5 is a schematic diagram of this potentiometer. Any potentiometer having a constant total resistance independent of the setting is suitable.

[4] C. L. Dawes, "A Course in Electrical Engineering," vol. 1, McGraw-Hill Book Company, New York, N. Y., 1937, p. 184.

As the normal plate current drawn by the voltmeter is only 0.001 ampere, a simple power supply using a 1-V rectifier tube suffices. Two glow-discharge voltage-regulator tubes connected in series maintain the plate-supply voltage constant at 300 volts. The heater of the 6F5GT has enough "thermal inertia" to eliminate fluctuations of plate current due to rapid variations of heater voltage. It is desirable, however, to connect the vacuum-tube voltmeter to a small voltage-regulated transformer for maximum stability, where frequent variations of line voltage are encountered.

The meter is a 5-inch diameter, 0 to 1.5 milliammeter accurately calibrated at the 1-milliampere point. Two terminals connected in series with the milliammeter are provided on the panel of the vacuum-tube voltmeter in order that a highly accurate external meter, or other calibration device may be conveniently employed. It is to be noted that the use of an external meter has no effect on the calibration of the voltmeter.

The saturation value of the plate current with the

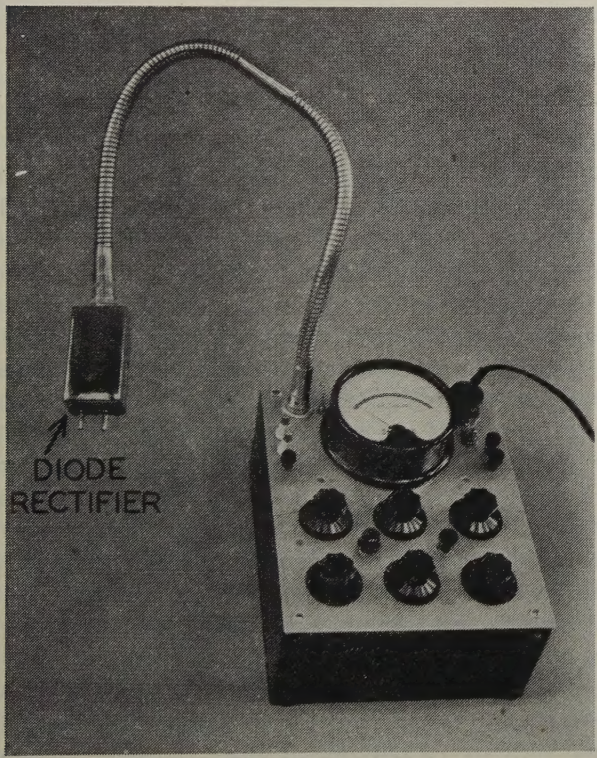


Fig. 4—Experimental model of the vacuum-tube voltmeter.

circuit constants and the plate-supply voltage employed is less than 3.0 milliamperes. This value of current is insufficient to cause severe damage to the meter on overload. If the grid becomes positive on large overloads, the milliammeter will read the sum of the plate and the grid currents. The resistance in the cathode circuit provides some degree of protection. If the normal grid current at the balance condition is negligible, the meter may be placed directly in the plate circuit. The milliammeter may be completely protected by connecting a 1-megohm resistance in series with the grid of the tube.

When this vacuum-tube voltmeter is used to measure the output voltage of a diode voltmeter connected to a resonant circuit, it will give a direct qualitative indication of resonance in the circuit. Resonance is indicated by a maximum increase in the cathode current for any particular setting of the Thomson-Varley

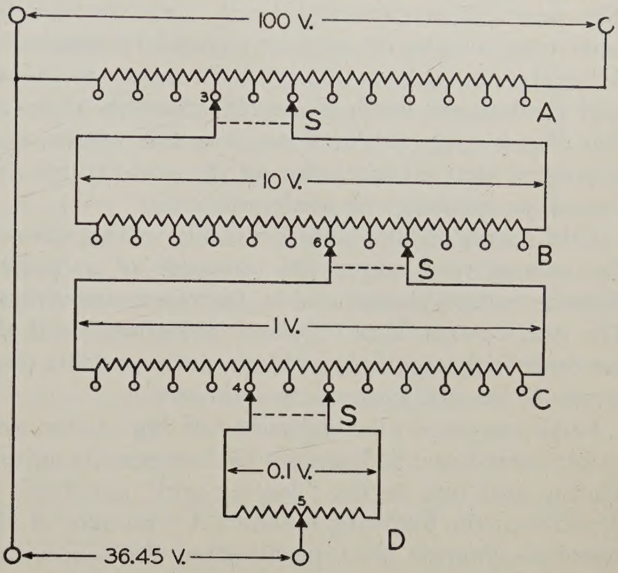


Fig. 5—100,000-ohm Thomson-Varley potentiometer.
A = eleven 10,000-ohm resistors
B = eleven 2000-ohm resistors
C = eleven 400-ohm resistors
D = calibrated 800-ohm potentiometer
S = two-gang, ten-position switches connected so as to bridge over two resistors in series as indicated.
With 100 volts applied to the potentiometer, the output is 36.45 volts for the switch positions indicated in the figure.

potentiometer. After resonance has been established, it is an easy matter to read the exact value of the voltage developed at the diode voltmeter output terminals by using the slide-back feature.

The stray voltage developed at the diode voltmeter output terminals with no input voltage is balanced out by an equivalent voltage developed across the variable cathode resistance $R_k - R_k'$. In order to effect this balance, the diode input terminals are shorted, the Thomson-Varley potentiometer is set at zero, and the cathode resistance $R_k - R_k'$ is adjusted for the reference cathode current of 0.001 ampere. The voltmeter is then ready for alternating-voltage measurements.

Grid Current

In a self-biasing circuit of the type used in the voltmeter, the grid is always negative with respect to the cathode at the reference balance condition. With this mode of operation, the input resistance of the tube is extremely high, as normally there is only a minute grid-circuit current[5,6] due to (1) electrons emitted by the cathode, (2) positive gas ions, (3) insulation leakage, (4) electrons emitted by the grid, and (5) positive metallic ions emitted by the cathode. The first three of these contribute most to the grid current.

[5] Institute of Radio Engineers, "Standards on Electronics," 1938, Section III-D, pp. 34–38.
[6] R. M. Bowie, "This matter of contact potential," Proc. I.R.E., vol. 24, pp. 1501–1513; November, 1936.

The grid emission current (4) is reduced in the practical voltmeter by operating the 6F5GT at a heater voltage of 5 volts instead of at the rated value of 6.3 volts. The insulation leakage current (3) is reduced by coating the glass envelope of the tube with ceresin wax and by carefully insulating the external grid circuit. The positive-ion current (2) will vary considerably from tube to tube. It may be reduced by employing a low plate-to-cathode potential. Variations in the net grid current are much more objectionable than the fact that a grid current flows. For this reason, it is important that all insulation in the grid-cathode circuit of the voltmeter be nonhygroscopic.

Grid current causes an appreciable voltage drop in the voltage read across the terminals of a high-resistance voltage source, and is, therefore, undesirable. The grid current is particularly objectionable if the internal resistance of the voltage source and the magnitude of the grid current are unknown.

In the vacuum-tube voltmeter of Fig. 3, the grid current is reduced to less than 10^{-9} ampere by adjusting the grid bias to the "floating-grid" potential[7] of the tube in the following manner: A resistance of 100 megohms shunted by a push-button short-circuiting switch is connected to the input terminals of the voltmeter. Then with the calibrated potentiometer set at zero, the cathode resistor $R_k - R_k'$ is adjusted for the reference value of plate current of 1 milliampere. If the plate current changes when the high resistance is short-circuited, a new value of plate-supply voltage is selected by means of the resistance R_1 in Fig. 3, and the plate current is again adjusted to 1 milliampere. A combination of plate and bias voltages is readily found which gives no change, or a minimum change, of plate current when the 100-megohm resistance is short-circuited. In the event that there is a change in plate current when this resistance is short-circuited, the voltage drop developed across it is measured by readjusting the plate current to 1 milliampere by means of the calibrated potentiometer. The grid current is then calculated from the known value of the 100-megohm resistance and the voltage drop as read from the potentiometer setting. It is understood that these extreme precautions to reduce, or to calculate, the grid current are necessary only when a minute grid current may not be tolerated.

Calibration and Accuracy

The inherent over-all accuracy of the voltmeter is determined by the precision of design of the potentiometer, the accuracy of the calibration of the meter at a single point, and the stability of all circuit parameters during a measurement interval. It is important that the resistors comprising the potentiometer have a low temperature coefficient of resistance, and are free from aging effects. These resistors need dissipate a total of only one tenth of a watt when the voltmeter is balanced, and less than a watt under extremely unbalanced conditions. The temperature rise of the resistors is, therefore, negligibly small.

The vacuum-tube voltmeter is completely calibrated upon construction, if the 1-milliampere point is accurately marked on the milliammeter. This point may be checked at any time by connecting an accurate laboratory milliammeter to the calibration terminals marked B in Fig. 3. The completed vacuum-tube voltmeter may be precisely calibrated by checking it against a standard cell.

In using the vacuum-tube voltmeter for direct-voltage measurements, no difficulty is encountered in repeating readings to four significant figures for voltages in excess of 1 volt. As fewer dials of the potentiometer are brought into use when voltages lower than 1 volt are read, the number of significant figures obtainable is necessarily reduced. It must be remembered, however, that the ultimate per cent accuracy of the experimental model of the vacuum-tube voltmeter is limited by the accuracy of the resistors comprising the potentiometer.

Conclusions

Standard design procedure may be employed to construct a low-voltage, battery-operated, vacuum-tube voltmeter using the circuit described in this paper with reduced element potentials to insure a low grid current. Long battery life is insured by the low plate current required. If great accuracy of voltage measurement is desired, it is necessary only to match the individual resistor units composing the calibrated potentiometer to a good degree of precision. The absolute resistance of the potentiometer is then of secondary importance, since the voltage drop across the potentiometer is readily standardized by comparison with a standard cell. In this case, the milliammeter is calibrated at a point determined by the voltage range to be covered by the voltmeter.

Although the vacuum-tube voltmeter described in this paper is not direct reading, this disadvantage is offset by the fact that great accuracy is obtained over the complete range of the instrument from 0 to 100 volts without employing a range-changing arrangement. It must be remembered that the accuracy of many vacuum-tube voltmeters is dependent on the calibration of a d'Arsonval-type meter with an accuracy specified in per cent of full scale. The accuracy of these meters at one-fourth scale, or lower, may be quite poor.

In addition to its utility as a general-purpose meter, the practical model of the voltmeter has proved particularly useful as a reference standard in calibrating direct-current voltmeters and milliammeters. Not only is there little danger of injuring the meter on overload, but it also provides a precision of reading which is obtainable only in the more expensive types of laboratory standard instruments.

[7] R. H. Cherry, "The Measurement of Direct Potentials Originating in Circuits of High Resistance," Bulletin printed by Leeds and Northrup Company, Philadelphia, Pennsylvania, 1938, Reprint E-96R(1) 180–738.

New Magnetic Materials*

W. E. RUDER†, NONMEMBER, I.R.E.

Summary—*With the rapid growth of the radio and communication industry, a need for magnetic materials having special properties for this particular application has developed. A number of nickel-iron alloys, such as permalloy, nicaloi, Mu Metal, and variations of these have found wide application as they all have the common property of high permeability at relatively low inductions. Where high resistivity also is desired, additional alloying elements, such as chromium and molybdenum, have been added. Complete freedom from strain, either mechanical or chemical, is necessary for good magnetic quality, and the strain set up by magnetization can be compensated for in many cases by heat treatment in a magnetic field. Silicon-iron alloys and some of the nickel-iron alloys can be very much improved by a combination of cold-rolling and heat treatment which induces a high degree of preferred orientation. This cold-rolled strip has found wide application in various types of electrical apparatus.*

Permanent-magnet alloys of the alnico type have been very greatly improved recently so that the external energy factor (BH_{max}) is now about three times what it was in the best alnico heretofore available. Comparative data on the different types of permanent-magnet steels and alloys are given, and the new material should find wide application in the radio field. Considerable saving in material and size and weight of apparatus can be made by the application of these outstanding recent developments in magnetic materials provided suitable changes in design are made to allow for the most economical use.

WE ARE living in an age of custom-built alloys where every mechanical device has some metal or alloy which is best suited to each of its parts and formed or treated to give optimum service. Parts for magnetic circuits are no exception. Since Stanley built his first transformer of "tintype" steel, probably because it was thin and had a varnished surface, the metallurgists have been kept busy to find the materials and treatments to meet the rapidly expanding and increasingly exacting requirements of the electrical engineer.

For some years the development of magnetic materials centered around low losses and high permeability at relatively high densities, for the engineer was concerned with increased efficiency and the decrease in size of his apparatus. To this end silicon-iron alloys were developed to a high degree and losses have been reduced from 1.20 to 0.35 watt per pound, 60 cycles, at 10,000 lines of force, through careful control of composition, structure, and heat treatment. Although many another has been tried, no completely satisfactory substitute for silicon as an alloying element has been found. The study of the nickel-iron alloys developed a most interesting range of magnetic and electrical properties, but the most interesting of these had a very high alloy content and relatively low magnetic saturation, so that they found only limited application for economic reasons.

With the rapid growth of the radio and communication industry, however, greater interest was shown in materials having high permeability in the low induction range, and the nickel-iron alloys came into their own. Permalloy, developed by the Bell Telephone Laboratories; the 50 per cent nickel-iron alloy now being made under a number of trade names; Mu Metal; and a number of other variations are now standard products, and permeabilities of a magnitude unheard of some years ago are readily obtainable. The number of magnetic alloys proposed to serve such special ends as high initial permeability, high resistivity, constant permeability, low loss at very high frequencies, and combinations of these are legion. For other than power applications, most of these are nickel-base alloys. The function of other alloying elements is principally to increase resistivity. There are certain underlying principles which govern the magnetic quality of all soft magnetic materials, most of which have to do with the control of internal strain.

The practice of careful annealing to remove mechanical strain is of long standing. There are, however, other sources of strain which are not relieved by temperature alone. Impurities, the presence of

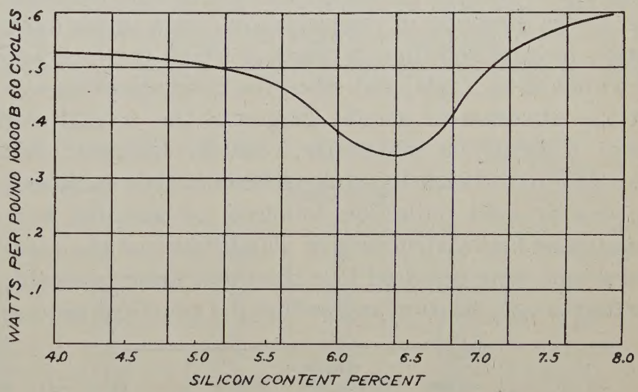


Fig. 1—Effect of silicon content on watt loss.

which disrupt the orderly arrangement of the atoms, set up strain, so freedom from dissolved impurities, usually nonmetallic, is of first importance. There are also strains set up by magnetization itself, the change in length called magnetostriction. The effect of grain size is probably also a manifestation of strain due to grain-boundary discontinuities, but the effect of crystal orientation is another matter.

In manufacturing high-grade transformer sheet these principles are applied to the greatest degree found to be commercially practicable. Silicon has the effect of increasing resistivity and reducing the solubility of the alloy for carbon, thereby increasing the magnetic quality at medium and low density, but it has the limiting effect of reducing the saturation induction. The highest quality that has been obtained in commercial silicon alloys is in a $6\frac{1}{4}$ per cent silicon sheet (see Figs. 1, 2, and 3), where high permeability and low loss (under 0.35 watt per pound, 60 cycles, 10,000 lines of force) result from the use of an alloy having practically zero magnetostriction. (78 per cent nickel permalloy is another example of such an

* Decimal classification: R282.3. Original manuscript received by the Institute, December 17, 1941. Presented, Rochester Fall Meeting, Rochester, New York, November 11, 1941.

† Research Laboratory, General Electric Company, Schenectady, New York.

alloy.) The alloy is quite brittle, although it can be rolled, punched, and sheared without difficulty.

For many years engineers have known that there is a preferred direction in magnetic sheets and that the properties are distinctly better in the rolling direction. Careful study of the magnetic properties of single crystals shows this to be due to the fact that the rolling develops a certain degree of preferred orientation of the crystals. The measurements made in this laboratory in 1917–1918 showed that silicon ferrite crystals oriented so that they presented a cube face perpendicular to the direction of magnetization were much more easily magnetized than crystals in which the face was oriented at an angle; and when the magnetization was in the direction of a cube diagonal, the crystal was most difficult to magnetize.[1] Smith, Garnett, and Randall[2] found that ferromagnetic materials subjected to drastic cold reduction followed by suitable heat treatment had a structure in which most of the crystal grains were oriented in a direction most favorable to easy magnetization and so found a practical process for producing a ferromagnetic material approaching in properties those of a single crystal. N. P. Goss,[3,4] working with the Allegheny Ludlum Steel Corporation on silicon iron alloys alone, found that the same general principle of cold reduction and suitable heat treatment produced a low-loss high-permeability strip and obtained patents on such a process.

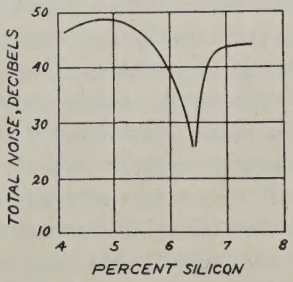


Fig. 2—Effect of silicon on noise.

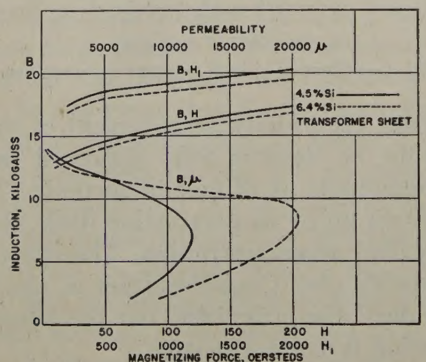


Fig. 3—Magnetization and permeability of normal and high silicon transformer sheets.

The commercial development of these processes has made available a high-grade silicon-alloy strip which, when applied so that it is magnetized in the direction of rolling, can be used at much higher densities than usually used with hot-rolled sheets. Fig. 4 shows a comparison of magnetization curves in the medium to high-induction range for hot-rolled sheet and cold-rolled strip with various degrees of preferred orientation. This new material has made possible radical changes in design[5] and improvements in efficiency of distribution and power transformers.

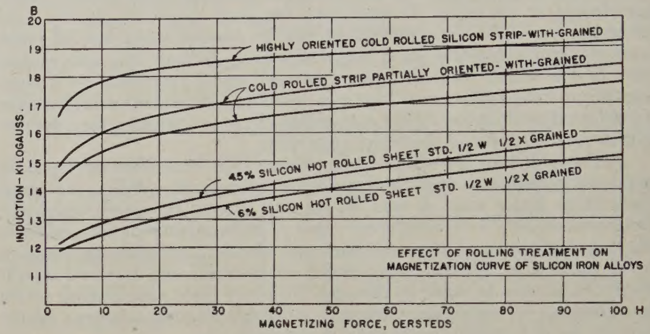


Fig. 4

In the low-density range, cold rolling and magnetic annealing may also be applied to advantage. The degree of improvement obtained varies with the alloy. An idea of the magnitude of these changes at low induction may be obtained from the figures given in Table I.

TABLE I

	Mu_{max}	B at $H=0.05$	Mu at $B=1000$	Mu at $B=100$
Silicon sheets—transformer grade	13200	165	5800	2950
Silicon sheets—magnetic anneal	20000	240	7950	4100
6 per cent silicon sheet	18000	140	5700	2500
6 per cent silicon—magnetic anneal	32100	220	8600	3450
Cold-rolled silicon strip	30500	700	6700	1600
Cold-rolled—magnetic anneal	42800	700	7150	1700
Nicaloi—hot-rolled	45500	1545	—	7400
Nicaloi—cold-rolled	94500	4600	44000	13100
Cold rolled + magnetic anneal	161000	7800	81800	29000
Allegheny Mu Metal	100500	3413	100000	74250
"65" permalloy—cold rolled	8500	160	6890	2830
"65" plus magnetic anneal	490000	9650	337500	105000

There has been a great advance in the field of permanent-magnet materials in the last ten years. This advance has opened many new applications as well as simplified many of the older applications of permanent magnets. Table II lists a number of these and their properties.

The best of the quench-hardening types have a maximum external energy value (BH_{max}) of about 1,000,000. Carbon-free magnet alloys were first developed along about 1930.[6–9] Alloys of this type

[1] W. E. Ruder, "Magnetization and crystal orientation," *Trans. Amer. Soc. for Steel Treating*, vol. 8, pp. 23–29; July, 1925.

[2] W. S. Smith, H. J. Garnett, and W. F. Randall, United States Patent No. 1,915,766.

[3] N. P. Goss, "Written discussion," *Trans. Amer. Soc. Metals*, vol. 22, pp. 1133–1139; December, 1934.

[4] N. P. Goss, "New development in electrical strip steels characterized by fine grain structure approaching the properties of a single crystal," *Trans. Amer. Soc. Metals*, vol. 23, pp. 511–531; June, 1935.

[5] E. D. Treanor, "The wound-core distribution transformer," *Elec. Eng.*, vol. 57, pp. 622–625; November, 1938.

[6] W. Kroll, French Patent No. 669,551, 1929.

[7] R. S. Dean, United States Patent No. 1,904,859, 1933.

[8] W. Köster, "Permanent magnets of the precipitation hardening type," *Stahl und Eisen*, vol. 53, pp. 849–856; August, 1933.

[9] W. Köster, "Mechanical and magnetic precipitation hardening of iron-cobalt-tungsten and iron-cobalt-molybdenum alloys," *Archiv. für Eisenh.*, vol. 6, pp. 17–23; July, 1932.

TABLE II

TYPICAL VALUES OF SOME PERMANENT-MAGNET ALLOYS

Composition (approximate)	H_c	B_r	External Energy B_dH_d max$\times 10^6$
Quench Hardening			
1% carbon steel	48	8600	0.18
6% tungsten steel	72	10000	0.32
36% cobalt, 4% tungsten, 5% chromium	240	9600	0.98
Dispersion Hardening			
12% cobalt, 17% molybdenum, 0.1% carbon	250	10500	1.20
Alnico I	440	7300	1.40
Alnico II	560	7350	1.65
Alnico III	475	6900	1.38
Alnico IV	730	5300	1.30
Alnico V	550	12500	4.5
New "KS"	785	7150	2.03
Superlattice			
76.7% platinum, 23.3% cobalt	2650	4530	3.77
78% platinum, 22% iron	1750	5380	3.07
86% silver, 9% manganese, 5% aluminum	6980	5550	1.56
Oxide Magnets			
16 CoO, 34 Fe_2O_3, 50 Fe_3O_4	580	4050	1.34
16 CoO, 34 Fe_2O_3, 50 Fe_3O_4	915	2170	—
Cold-Worked			
60% copper, 20% nickel, 20% iron	513	5350	1.26
41% copper, 24% nickel, 35% cobalt	444	5325	0.99

According to Legg,[12] the theoretical limit for H_r is about 23,000. The closest approach to this was reported by Nesbitt and Kelsall.[13] They found a B_r value of 15,000 gausses in an iron-vanadium-cobalt alloy.

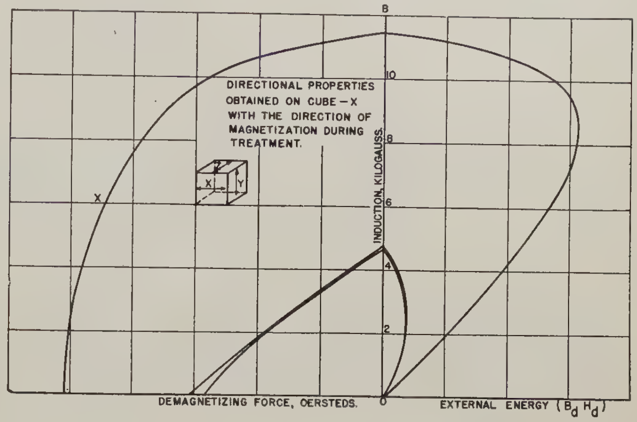


Fig. 6—Directional properties of alnico V.

which have found the widest application are in the nickel-aluminum-iron group, which is now known in this country and England as alnico.[10,11]

The general requirement for a good permanent-magnet material is a hysteresis loop of large area. The area of this loop is largely determined by the value of the residual (B_r) and the coercive force (H_c). The shape of the demagnetization curve, however, is also important as it determines the density at which a permanent magnet can be used with greatest efficiency.

Legg also places the theoretical limit for H_c at 23,000 oersteds. The nearest approach to this is a value of about 7000 oersteds for aluminum-manganese-silver.[14] It has now become customary to designate the quality of the magnet in terms of its maximum external energy, or BH_{max} value. As shown in the table, this has been increased from 1,000,000 in Honda's cobalt magnet steel to 1,800,000 in alnico II, to 2,000,000 in the "new KS," and to 3,770,000 in the platinum-cobalt alloy. Oliver and Shedden[15] improved the B_r value and the BH_{max} some 20 per cent by subjecting a nickel-aluminum-cobalt-copper permanent-magnet alloy to a strong magnetic field at high temperature. Jonas and van Embden,[16] working with this type of alloy, but with considerably higher cobalt content and in some cases with small copper and titanium additions, produced a remarkable increase in BH_{max}.

TABLE III

Composition in per cent					Usual thermal treatment			Same treatment but cooled in a magnetic field			Cooling rate 1200 degrees to 600 degrees centigrade
Nickel	Aluminum	Cobalt	Copper	Titanium	BH_{max} $\times 10^6$	H_c	B_r	BH_{max} $\times 10^6$	H_c	B_r	°C/second
16	8.5	23	—	—	1.22	348	9050	3.45	492	126500	1
13.5	8.0	24	1.5	—	1.32	370	9450	3.77	505	13100	2.6
13.5	8.0	24	3.0	—	1.68	535	8300	4.78	600	13350	1.8
16	7.8	25	—	2.8	1.60	604	7600	3.06	640	10000	4
14	7.1	24	3.0	2.4	1.72	594	7900	3.78	660	11050	4
14	7.5	20	6.5	1.8	1.65	620	7350	3.25	676	9825	4.3
16.5	8.1	20	1.1	2.3	1.82	640	8150	3.12	685	10200	4

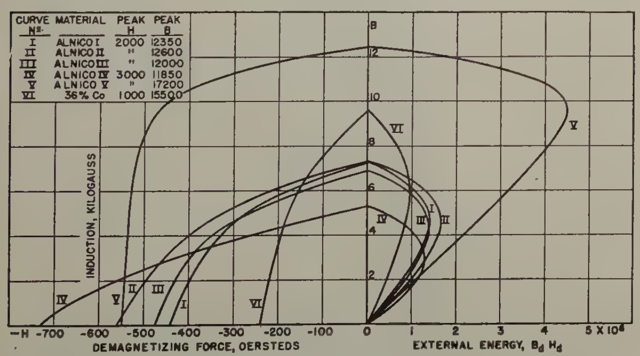


Fig. 5—Demagnetization and energy curves of alnico type of permanent magnets.

[10] W. E. Ruder, United States Patent No. 1,947,274; February, 1934.

[11] T. Mishima, United States Patent No. 2,027,994, 1936.

[12] V. E. Legg, "Survey of magnetic materials and applications in the telephone system," *Bell Sys. Tech. Jour.*, vol. 18, pp. 438–464; July, 1939.

[13] E. A. Nesbitt and G. A. Kelsall, "Vicalloy, a new permanent magnet material," *Phy. Rev.*, vol. 58, p. 203; July 15, 1940.

[14] H. H. Potter, "Some magnetic alloys and their properties," *Phil. Mag.*, vol. 12, pp. 255–264; August, 1931.

[15] D. A. Oliver and J. W. Shedden, "Cooling of permanent magnet alloys in a constant magnetic field," *Nature*, vol. 142, p. 209; July 30, 1938.

[16] B. Jonas and H. J. Meerkamp van Embden, "New kinds of steel of high magnetic power," *Philips Tech. Rev.*, vol. 6, pp. 8–11; January, 1941.

Details of this process and composition of the high-cobalt alloys are given in the British patent specification No. 522,731, from which Table III on page 439 is taken.

The product, shown in Fig. 5 as alnico V, has strongly directional properties, i.e., it must be magnetized for use in the direction of the field applied during the heat treatment. Fig. 6 shows the difference in properties in different directions.

Like all magnetic materials, optimum properties are obtained only with the most careful control of composition and heat treatment. It is possible with the right conditions to obtain values of BH_{max} as high as 6,000,000, with B_r up to 13,500, and H_c up to 700. The high cobalt content makes the cost per pound relatively high, but the high energy content with the resultant saving in size brings the cost per unit of available energy to a figure quite comparable with the other alnico compositions.

The application of these two outstanding recent developments in magnetic materials calls for radical changes in design in each case if their use is to be justified, but in all applications where such changes are made, considerable savings in material and reductions in size and weight of apparatus can be made.

Analysis, Synthesis, and Evaluation of the Transient Response of Television Apparatus*

A. V. BEDFORD†, ASSOCIATE, I.R.E., AND G. L. FREDENDALL†, ASSOCIATE, I.R.E.

Summary—*The sharpness of detail in a television picture is directly dependent upon the capability of the transmitter for the transmission of abrupt changes in picture half tone. A suitable test signal is a square wave of sufficiently long period.*

Rules are deduced for the evaluation of the subjective sharpness to be expected in transmitted pictures and may be applied when the square-wave response of the transmitting apparatus is known.

Rapid chart methods have been devised for (1) the analysis of a square-wave output into sine-wave amplitude and phase response and (2) the synthesis of a square-wave response from a given set of amplitude and phase characteristics. Analysis furnishes an immediate solution to the familiar but troublesome problem of finding the sine-wave characteristics of television apparatus.

The four aspects of the application of square waves to television, i.e., measurement, analysis, synthesis, and evaluation, are presented as a basis for a unified and complete technique.

The authors hope that this paper will be a contribution to the general problem of working out electrical specifications for television transmitters and other television apparatus, giving information regarding the steepness of rise and the amplitude of overswing of the square-wave response.

I. Introduction

As a result of the scanning process, the sharpness of detail in a television picture is directly dependent on the capability of the transmitter and receiver for the faithful transmission of signals arising from abrupt changes in picture half tone along the scanning line. Recognition of the validity of the Heaviside unit voltage, the electrical equivalent of an abrupt change in half tone, as a test signal, was accorded early in the art. Notwithstanding, the preponderance of emphasis has been placed upon the sine-wave characteristics of television apparatus, that is, upon the amplitude- and phase-versus-frequency characteristics.

It has long been known that the response of a linear signal-transmitting system to a Heaviside unit voltage contains all the information necessary to determine both the phase-frequency and the amplitude-frequency characteristics. Conversely, the two frequency characteristics determine uniquely the response to a unit voltage (and incidentally the response to any other transient input wave). It is also known that the response to a single abrupt rise in a *repeating* square wave of sufficiently long period is essentially the same as the response to a Heaviside unit voltage insofar as that part of the transient response due to high frequencies is concerned.

In view of this implicit relationship between the sine-wave and the transient-response[1] characteristics of electrical circuits, it is surprising that the testing and specifying of high-frequency fidelity of both audio and video apparatus by the response to a square-wave has not become more common.[2]

Several circumstances have impeded rapid growth in the use of square waves to determine the high-frequency response. In the first place, the well-established sine-wave methods, which were developed for audio work, have the advantages of precedence and well-developed techniques for measurement, recording and plotting data, diagnosing imperfection, and comparing performances. Second, the lack of suitable oscillographic apparatus for accurately indicating the instantaneous response as a function of time has been a large contributing factor. A recently developed square-wave oscillograph[3] and square-wave generator provide a solution to this problem.

* Decimal classification: R583. Original manuscript received by the Institute, February 19, 1942. Presented, Summer Convention, Detroit, Michigan, June 25, 1941.

† RCA Manufacturing Company, Inc., Camden, N. J.

[1] In this paper, the term *transient response* will be used as the equivalent of the expression *response to a Heaviside unit voltage*.

[2] The use of a 60-cycle square-wave generator and an oscillograph for investigation of the behavior of a television system at *low frequencies* of the order of the field scanning rate is well established. In these measurements, performance is judged by inspection of the "tilted" output wave and harmonic analysis of the wave is usually not desired. In general, a television system will have uniform response over a frequency range of many octaves in the region between the so-called "low-frequency" end and the "high-frequency" end, so that fidelity measurements at the two ends of the spectrum may be considered separately. This paper will be concerned only with the high-frequency end.

[3] R. D. Bell, A. V. Bedford, and H. N. Kozanowski, "A portable high-frequency square-wave oscillograph for television," Proc. I.R.E., this issue, pp. 458–464.

The laboriousness of classical methods for translating the results of sine-wave measurements into transient response and vice versa, has tended improperly to make the two test methods seem unrelated and competitive instead of complementary. Furthermore, there has been no satisfactory means for evaluating *numerically* the fidelity of a particular piece of apparatus or transmitting system from the transient response. In television applications it has been recognized that the "mean" steepness of rise of the transient-response wave is a measure of the sharpness of the picture which the system could transmit. Still there is no general agreement on a method of measurement and calculation of a mean value of steepness which is a consistent and an accurate indication of the picture sharpness for a wave having nonuniform steepness during the time of rise. The presence of overswing in the transient response makes evaluation even more difficult because of the effect on the visual sharpness of the picture and because of the introduction of spurious effects in the picture.

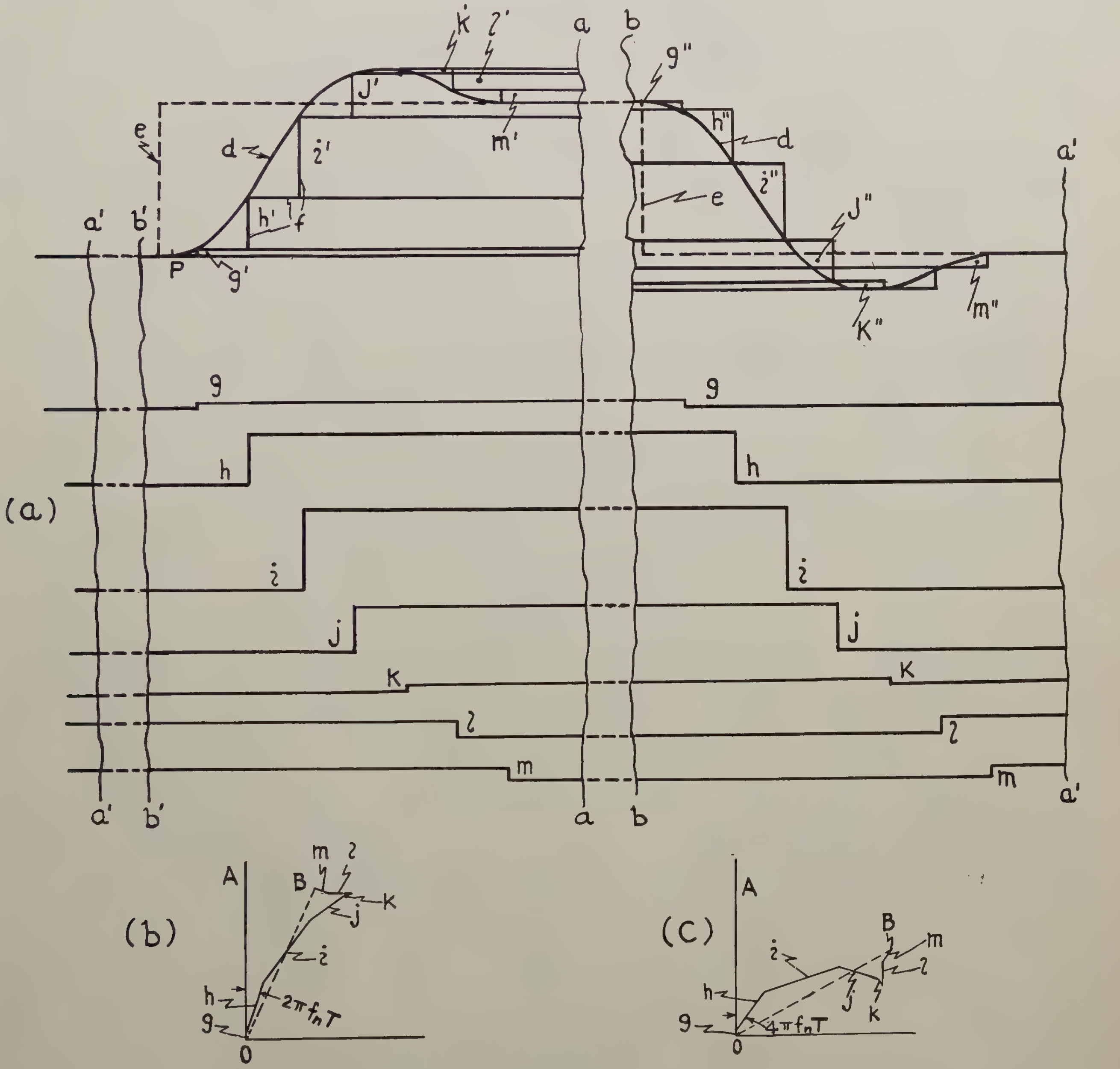


Fig. 1—(a) Approximation of a square-wave response d by a stepped wave f.
(b) Vector addition of components of frequency f_n in waves g, h, i, etc.
(c) Vector addition of components of frequency $2f_n$ in waves g, h, i, etc.

Accordingly, for the purpose of simplifying the passage between sine-wave response and transient response and interpreting the latter, we present below: (1) a graphical chart method for analyzing the response of a system to a square-wave input signal to obtain the sine-wave phase and amplitude characteristics; (2) a graphical chart method for synthesizing the response to a square wave from the sine-wave phase and amplitude characteristics; (3) a method for

evaluating the mean steepness of a transient-response wave in terms of the *width of blur* produced in a television image by a wave which is similar in its visual effect and which has a *linear* change from one level to another; and (4) suggestions for the supplementing of sine-wave measurements by transient measurements.

period of e is not critical but must be taken long enough to insure that wave d has subsided to a substantially constant level during the latter part of each half cycle. Also the true time relation between waves d and e under test conditions has no significance in the present analysis and need not be known.

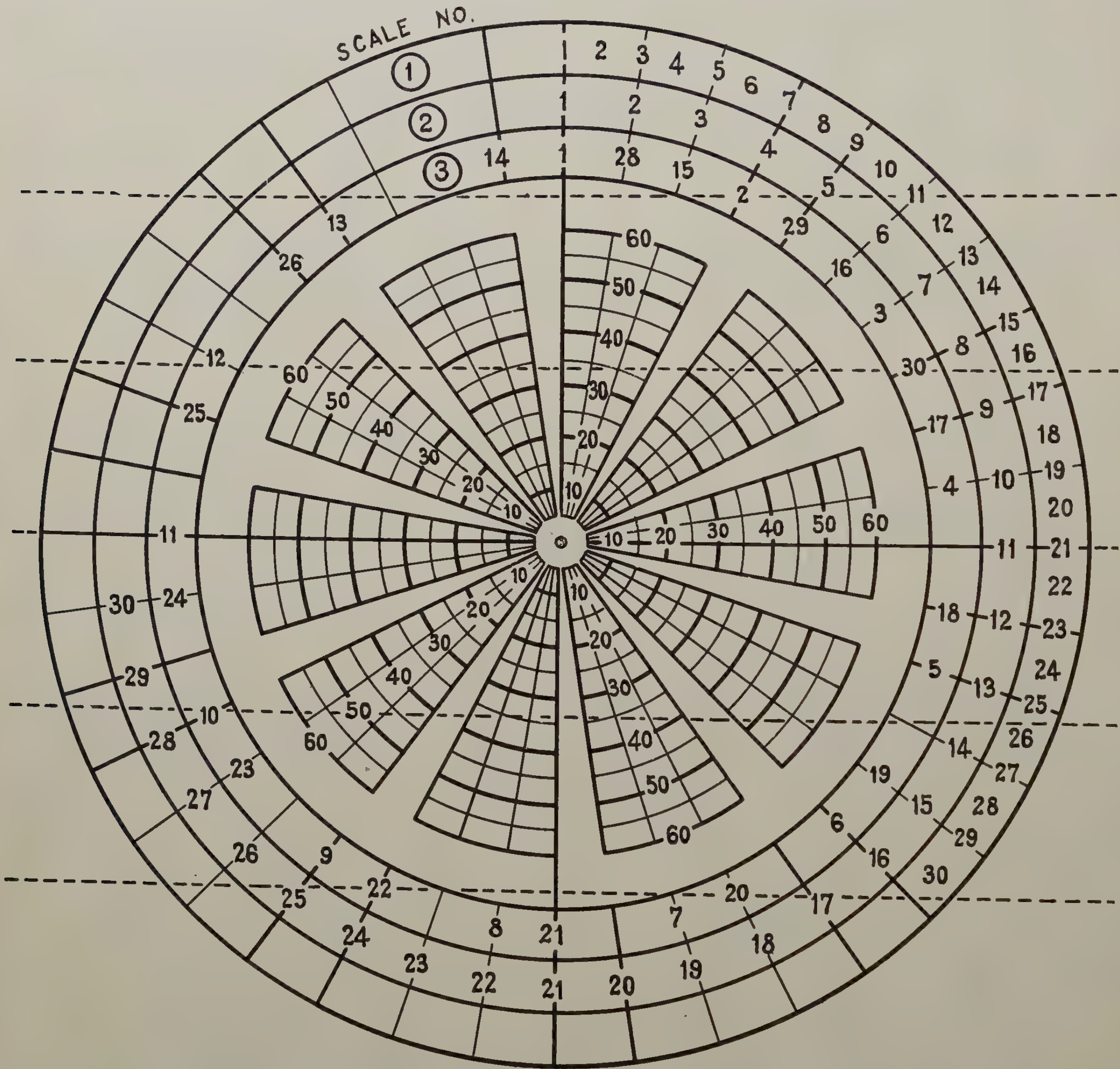


Fig. 2—Chart for square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
1	0.25 Mc	0.375 Mc
2	0.5 Mc	0.75 Mc
3	1.5 Mc	2.25 Mc

II. Analysis of Square-Wave Response into Phase-Frequency and Amplitude-Frequency Characteristics

The analysis employs several permanent charts, (Figs. 2 to 5), the construction and use of which may be simply explained by reference to Fig. 1.

Fig. 1 (a) shows the repeating square wave e applied to the apparatus under test. Wave d is the response measured at the output terminals. The fundamental

A basic hypothesis upon which the analysis of wave d rests is that a stepped wave f may be drawn which approximates in harmonic content that of wave d. It is clear from inspection that the waveform of f may be caused to approach that of e as closely as desired by taking the steps sufficiently small. Since wave f in the regions a to b and a' to b' is of uniform amplitude, the lengths of the time intervals from a to b and a' to b' have no bearing on the shape of the transient portion

of the waves. Hence, the fundamental period of wave f is unimportant provided that the value is great enough so that a part of f is uniform after each transition. The upward transition of wave f is the same as the downward transition except for inversion. Hence, the rectangles g'', h'', i'', etc., may be considered as relative to the same harmonic in the input wave e, is indicated by the sum of a group of vectors having amplitudes proportional to the steps g', h', i', etc., and angular positions corresponding to the different delays of the square waves g, h, i, etc. No cognizance need be taken of the fact that the harmonics of a square wave

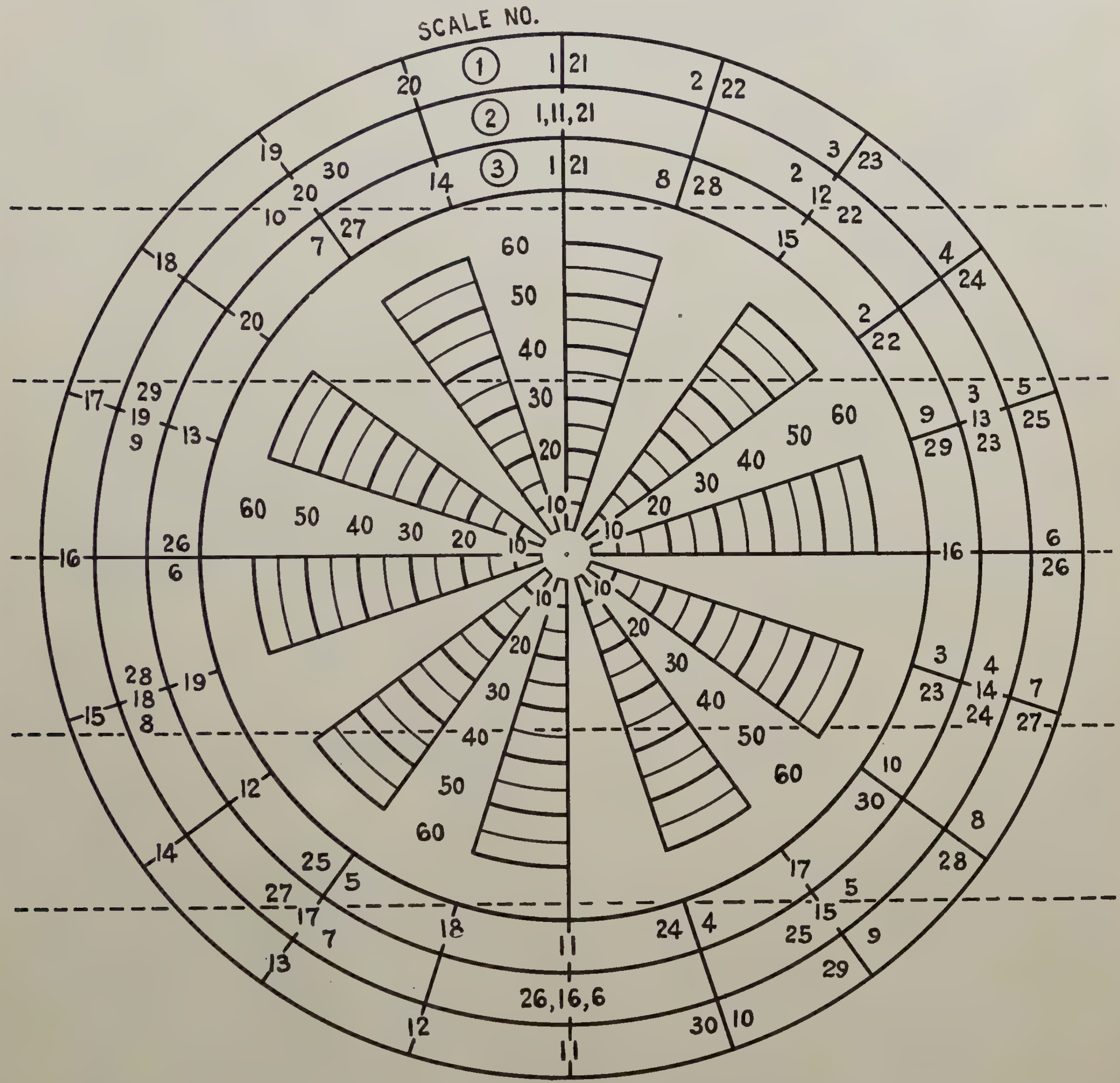


Fig. 3—Chart of square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
①	1 Mc	1.5 Mc
②	2 Mc	3 Mc
③	3 Mc	4.5 Mc

continuations of rectangles g', h', i', etc. It follows that wave f has the components g, h, i, etc.

Each of these square-wave components is identical in shape to the input wave e and hence contains all harmonics in the same proportion as the input wave. Each of the square waves, however, has a different delay and hence, the harmonics of the various square waves occur in different phase relations. Therefore, the magnitude of each harmonic in the stepped wave, vary in amplitude inversely as their frequencies. Inasmuch as the fundamental frequency of the input wave may be allowed to approach zero (such that wave e becomes a Heaviside unit voltage), any reference to discrete harmonics may be dropped and the response of the apparatus approximated at any frequency to a degree of accuracy which depends on the fineness of the steps in wave f.

Fig. 1(b) shows the vector addition involved in

finding the response for the frequency f_n. Each vector component has the same angular position, namely, $2\pi f_n T$, with respect to the preceding component since the waves g, h, i, etc. correspond to points on the real response curve taken at equal time intervals T. Waves l and m are negative; hence, the vectors l and m are

A set of analysis charts, Figs. 2, 3, 4, and 5, were designed in order to reduce to a minimum the labor involved in performing the vectorial additions in Figs. 1(b) and (c). Essentially, each chart serves as a protractor and a linear radial scale for use in locating the end points of the vectors. Two sets of time intervals,

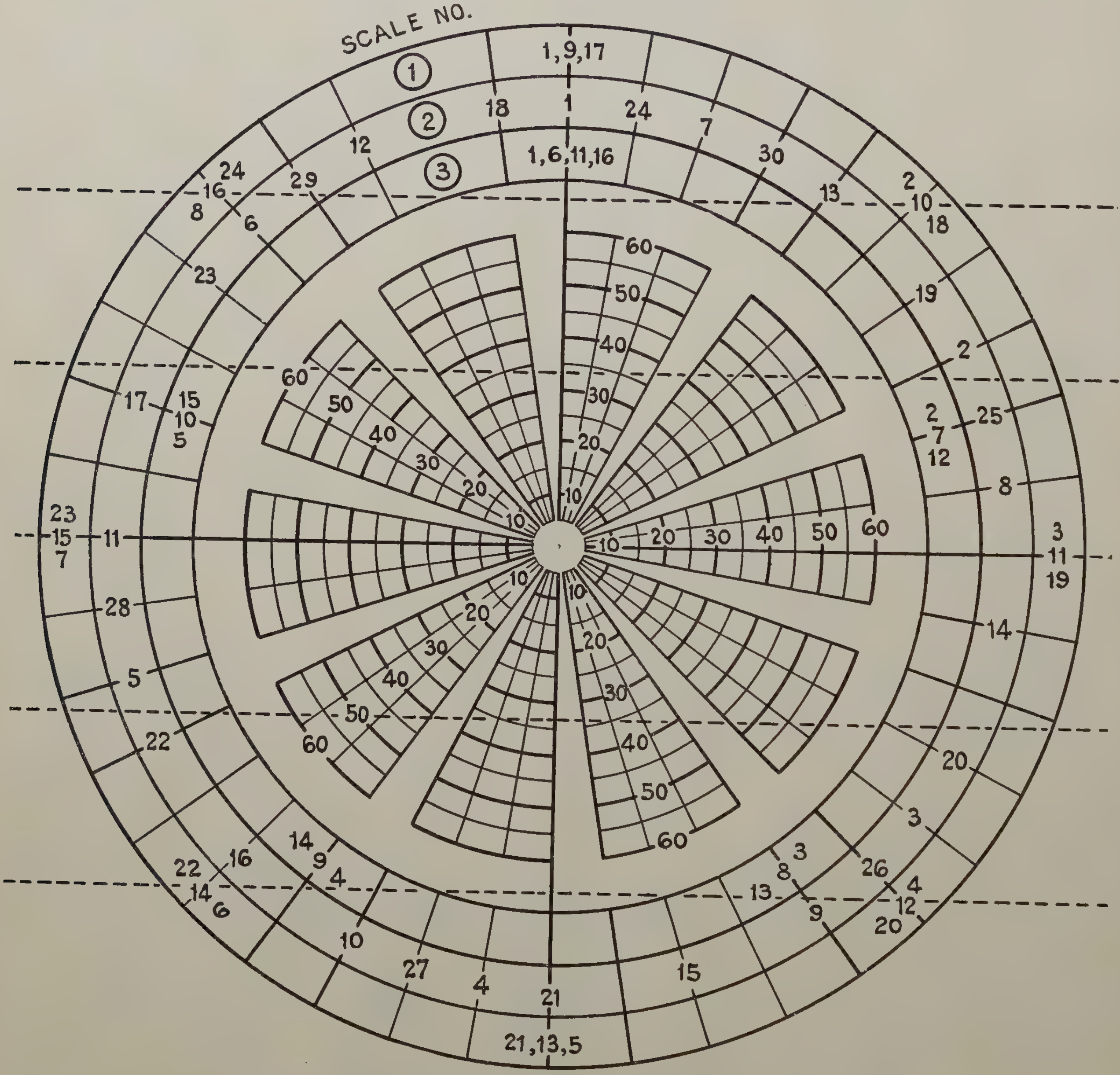


Fig. 4—Chart for square-wave analysis.

Scale	20-Mc Dots	30-Mc Dots
1	2.5 Mc	3.75 Mc
2	3.5 Mc	5.25 Mc
3	4.0 Mc	6.0 Mc

negative. The length of the vectorial sum OB gives the relative amplitude response of the apparatus tested and the angle AOB is the relative phase shift for the sine wave f_n. Fig. 1(c) is drawn for a frequency equal to $2f_n$. The angle between successive vectors is $2\pi 2f_n T$. As in Fig. 1(b), OB is the amplitude response at the frequency $2f_n$ and the angle AOB is the relative phase shift.

1/20 and 1/30 microsecond, (referred to as 20- and 30-megacycle dots) between successive components g, h, i, etc., appear to be adequate for television applications. The basis for a choice of one of the two sets depends upon the degree of accuracy desired. This aspect is discussed later. Components (i.e. readings from the transient-response wave d), are numbered to correspond to radial lines on the charts. The angle between

consecutively numbered radial lines is the angle used in the construction of Figs. 1(b) and (c) (e.g., on Fig. 3, the angle in scale 1 for the solution of the response at 1 megacycle is 18 degrees.) The component vectors g, h,

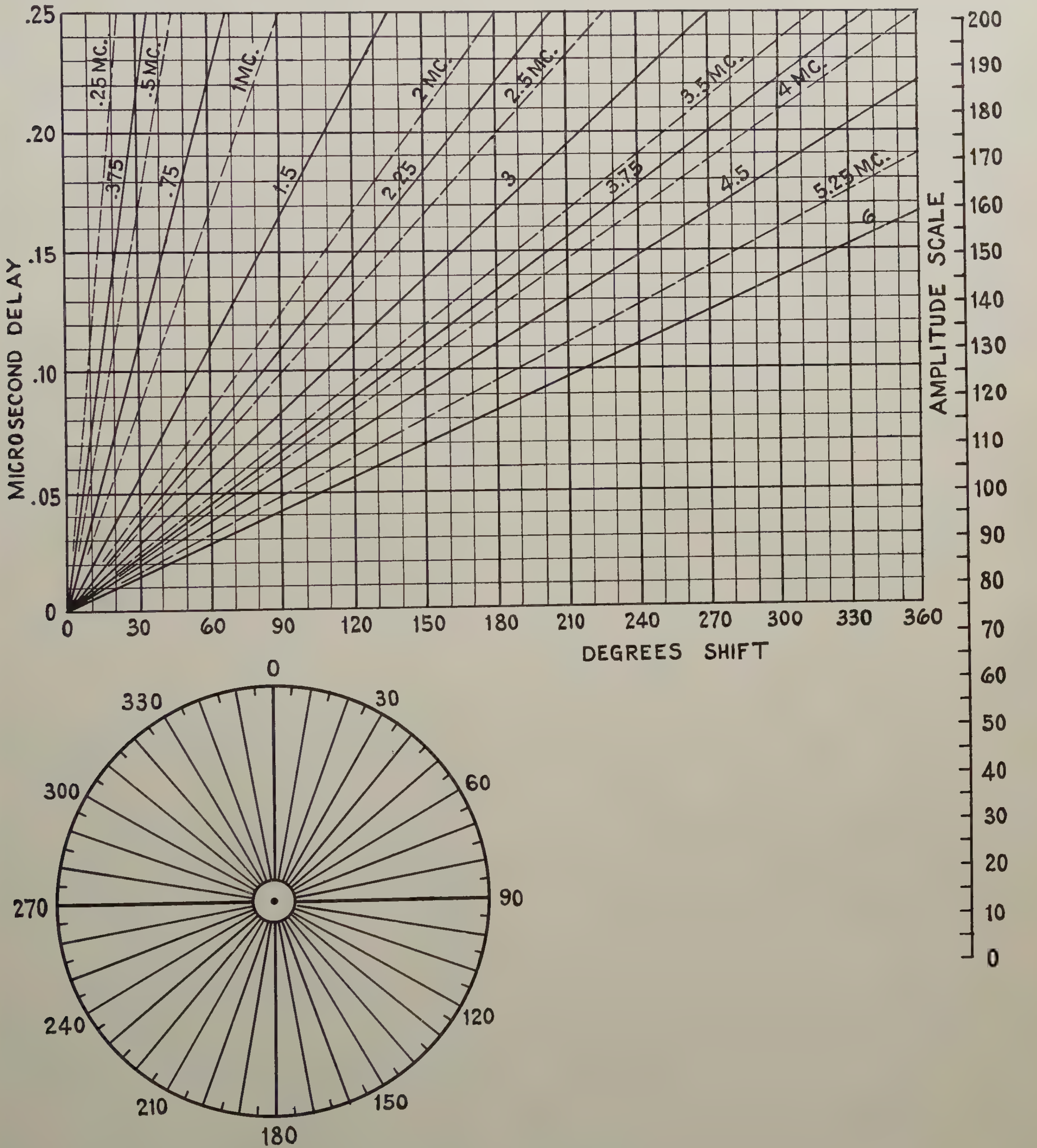


Fig. 5—Chart for square-wave analysis.

i, etc., lie along the radial lines and the vectors are added as in Fig. 1(b) by manipulation of a sheet of semitransparent paper. Detailed directions for the operation of the charts are given below. The charts and directions are drawn or printed preferably on cardboard or stiff paper.

A. Instructions for Using Figs. 2, 3, 4 and 5 for Analysis of Square-Wave Response

Using a square-wave generator and an oscillograph or other means, obtain per cent voltage readings at 1/20-microsecond intervals (corresponding to oscilloscopic readings with 20-megacycle "dots" for timing) along the transient wave (Fig. 6) such as shown in column (a) of Table I. (If more accurate analysis is desired, use readings at 1/30-microsecond intervals).

Readings should begin at the zero-voltage level

before the beginning of the transient and end at a 100 per cent point where the voltage becomes uniform after the transient rise has been completed and substantial rest attained. Wave d of Fig. 1 is a typical plot of such a wave but plotting is not necessary for the purpose of the analysis. Compile column (b) by taking the differences in adjacent readings in column (a). These numbers represent the amplitudes of the increments in the "stairstep" wave f. Number the increments consecutively as in column (c).

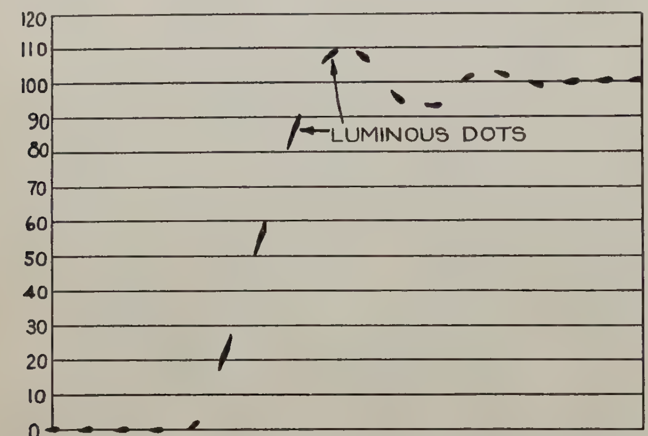


Fig. 6—Illustrative example of a square-wave response reproduced by an oscillograph showing readings at 0.05-microsecond intervals (20-megacycle dots).

Draw a horizontal x axis and a vertical y axis on a sheet of sufficiently transparent tissue paper. Place the tissue paper on Fig. 4, for example, with the origin of the x-y system at the center of the chart. (It will be noted that Figs. 2, 3, and 4 are for different frequencies.) Make a pencil dot on the tissue on the radial line marked "1" in circular scale (1) at a distance from the center of the chart equal to increment No. 1 in column (b). Move the tissue so that this dot coincides with the center of the chart, keeping the x axis parallel to the horizontal dotted lines on the chart. Make a second dot on the radial line marked "2" in circular scale (1) at a radius equal to increment No. 2. Move the tissue until the second dot is at the center of the chart and repeat the procedure for the remaining increments.

TABLE I

(a)	(b)	(c)	(a)	(b)	(c)
0	—	—	95	−13	7
2	2	1	93	− 2	8
22	20	2	101	8	9
56	34	3	102	1	10
87	31	4	99	− 3	11
108	21	5	100	1	12
108	0	6	100	0	13

Positive increments are plotted in the radial direction *toward* the increment number in circular scale (1) while negative increments are plotted in the opposite direction. Dots may be numbered to avoid errors.

Draw a vector from the x-y origin to the final dot located. Place the tissue on Fig. 5 and read the length of the vector on the "amplitude scale." This length is the 2.5-megacycle amplitude response in per cent. Read the angle between the y axis and the final vector by using the protractor in Fig. 5. This angle[4] is the phase lag of the 2.5-megacycle component of wave (d) relative to point P in Fig. 1 where P is 1/2 a reading-interval before the second reading in column (a). The phase angles obtained for the various frequency components are correct relative to one another but the absolute time delay through the system is not obtained. The phase angles may be converted to time delay by means of the delay graph of Fig. 5. In the example given above, the amplitude response at 2.5 megacycles is 85 per cent. The relative phase angle is 103 degrees which corresponds to a time delay of 0.114 microsecond. Obtain the response at other frequencies in a similar manner by using the increments in conjunction with the other scales of Fig. 2 and the scales of other charts.

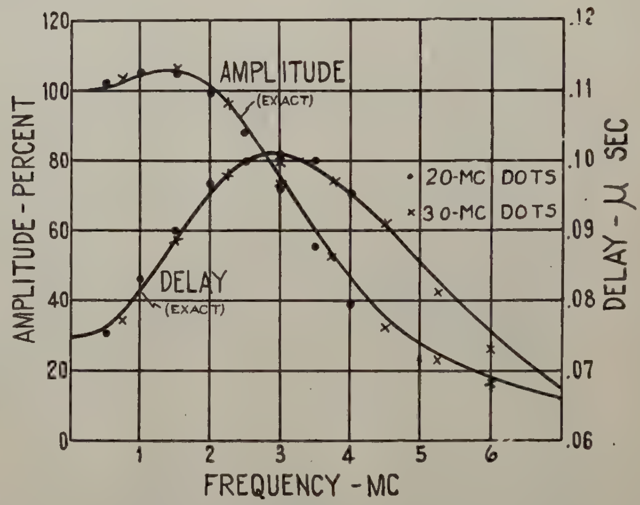


Fig. 7—Curves show exact amplitude and delay characteristics of a 2-stage compensated resistance-capacitance amplifier in which $K=\sqrt{2}$, $f_0=3$ megacycles. Amplitude and delay characteristics determined by chart analysis of Fig. 8 are shown by data points.

The charts may be used for other frequency ranges such as the audio range. In such applications, the time interval between successive readings on the transient-response wave will not usually correspond to 20- or 30-megacycle dots. However, the analysis charts shown in Figs. 2, 3, 4, and 5 may be employed as in the example given in Table I, when the appropriate multiplying factor for frequency is computed. This factor is equal to T_0/T_N where T_0 is $1/30\times10^{-6}$ or $1/20\times10^{-6}$ second depending upon the original dot frequency for which the charts were designed. T_N is the new time interval in seconds.

B. Examples of Accuracy of Chart Analysis in Specific Applications

In the instance of a compensated resistance-coupled amplifier, mathematical formulas are available for the exact calculations of the amplitude and delay

[4] The angle read between the y axis and the vector is the phase angle for the 2.5-megacycle response of the stepped wave f with respect to the time of the second reading of column (a). Wave f lags the true-response wave d by one half a reading interval.

characteristics and the response to a unit function.[5] Hence, some conception of the accuracy of chart analysis may be gained by observing the agreement of data thus determined with the exact characteristics. Fig. 7 contains such data based on the analysis of the unit-function-response wave shown in Fig. 8. Fig. 9 contains similar data based on the response wave of Fig. 10. Since the delay characteristics determined by graphical analysis have only relative significance, the data points corresponding to 20-megacycle-dot readings were shifted a constant amount (0.026 microsecond) in Figs. 7 and 9 so that the best correspondence with the mathematically determined delay curves was reached in order that the results of graphical and mathematical methods may be compared directly. A similar shift of 0.017 microsecond was made in the case of the 30-megacycle-dot data. The deviations of the data points in Figs. 7 and 9 from the exact curves represent the degree of approximation of analysis as applied to two specific cases. For the general case, no definite limits can be set up for the errror in sine-wave characteristics as determined by chart analysis. As in Fourier analysis,[6] the error depends upon the specific curve which is analyzed and upon the length of the time interval between readings, or in other terms, upon how well the curve is defined by the series of dots. A better approximation to the exact amplitude or delay characteristic at the higher frequencies is afforded by a solution based on 30-megacycle-dot readings rather than on 20-megacycle-dot readings.

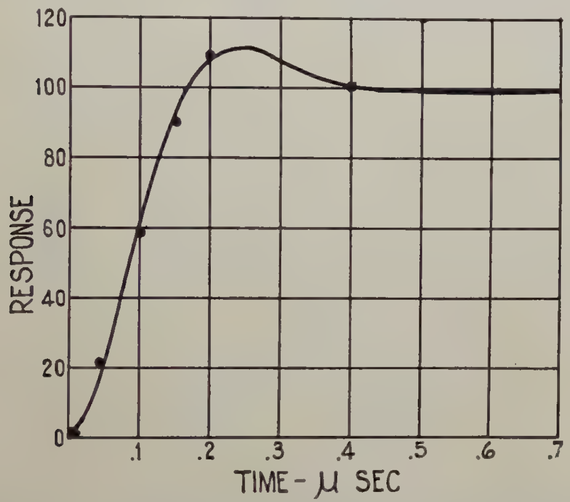


Fig. 8—Exact response to a unit function of a 2-stage compensated resistance-capacitance amplifier where $K=\sqrt{2}$ and $f_0=3$ megacycles. Points shown were obtained by synthesis from theoretical amplitude and delay characteristics in Fig. 7.

It is conceivable that as a consequence of strong components of very high frequency the transient-response wave could make violent excursions between adjacent readings such that a plot of the readings would not clearly define the wave even for the lower-frequency components. In such a case, inspection would show that analysis would be inaccurate. In general, it has been found that if the 20-megacycle timing dots trace out the square-wave response unmistakably, the 20-megacycle dots are sufficiently accurate for analysis out to 3.5 megacycles. Under the same conditions, 30-megacycle dots are adequate out to 5.25 megacycles.

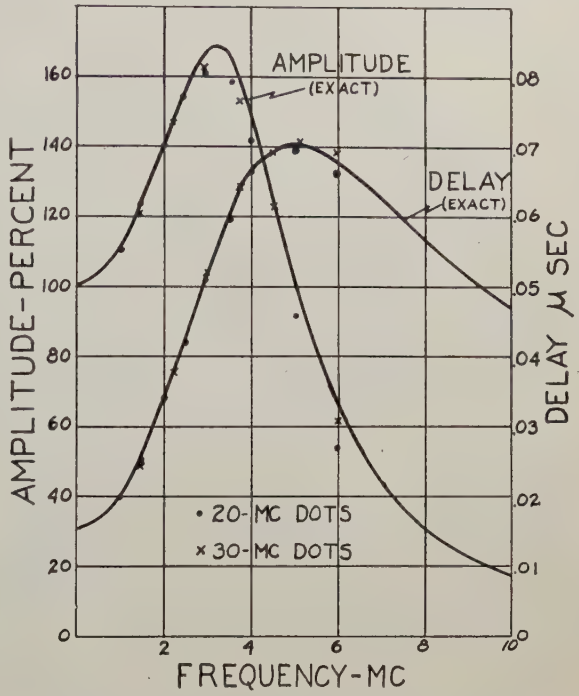


Fig. 9—Theoretical amplitude and delay characteristics of a 2-stage compensated resistance-capacitance amplifier in which $K=1.1$ and $f_0=4$ megacycles. Amplitude and delay characteristics determined by chart analysis of Fig. 10 are shown by data points.

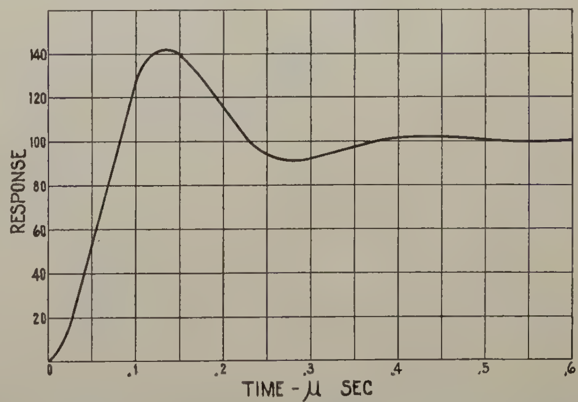


Fig. 10—Exact response to a unit function of a 2-stage compensated resistance-capacitance amplifier in which $K=1.1$ and $f_0=4$ megacycles.

III. Synthesis of Square-Wave Response from the Sine-Wave Characteristics

A closed mathematical formula[7] is available for the calculation of the response of a linear electrical system

[5] Appendix II.

[6] The amplitude characteristic of a circuit, as determined graphically from the square-wave response, may be converted into the amplitude characteristic corresponding to Fourier analysis of the response of the same circuit to a square *pulse*. This conversion is discussed in Appendix III.

[7] A. V. Bedford and G. L. Fredendall, "Transient response of multistage video-frequency amplifiers," Proc. I.R.E., vol. 27, pp. 277–285; April, 1939.

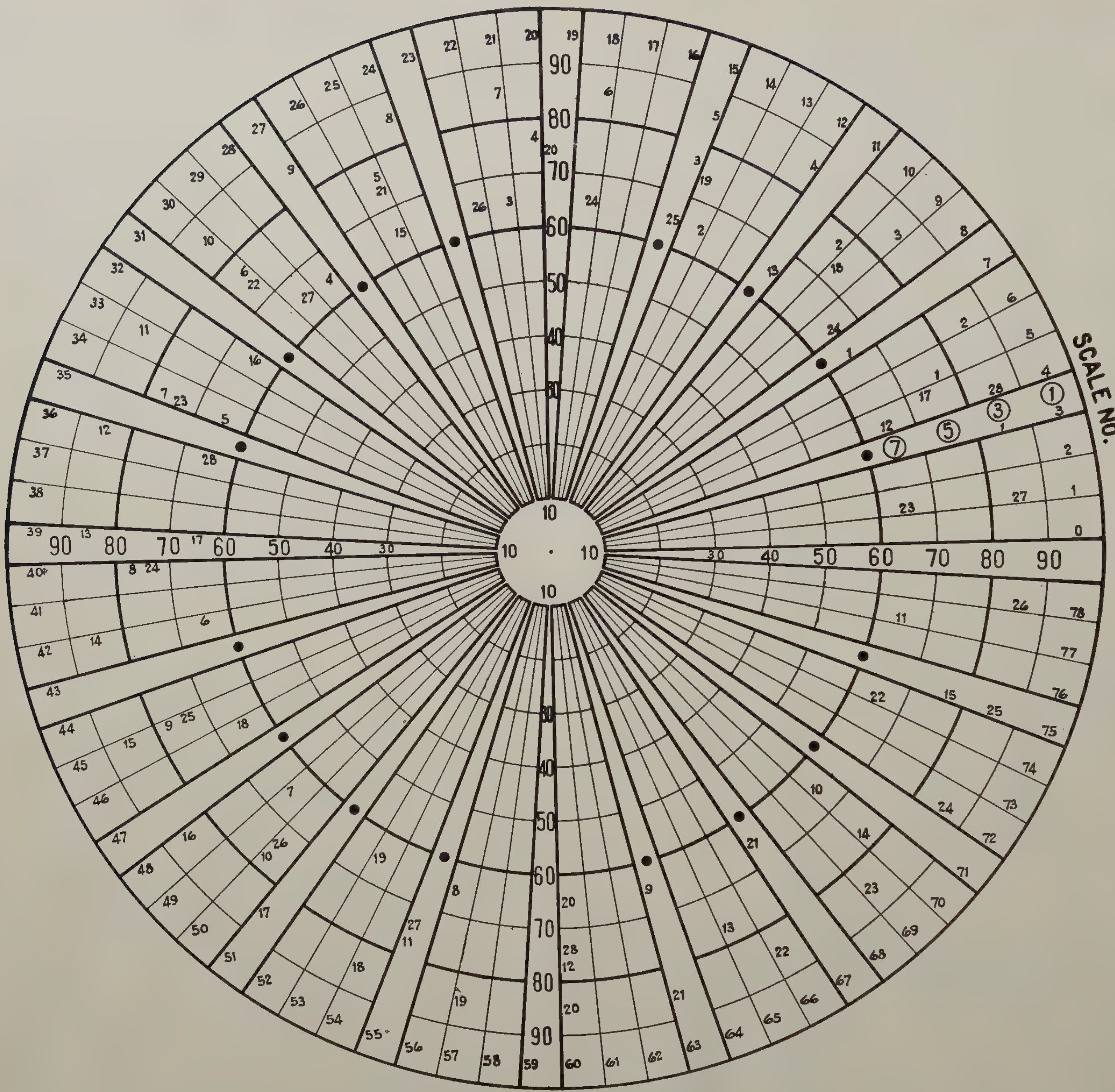


Fig. 11—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
1	0.25 Mc
3	0.75 Mc
5	1.25 Mc
7	1.75 Mc

to a unit function from the sine-wave amplitude and phase characteristics. When a numerical answer is sought, however, the formula is rarely practicable. As pointed out earlier, the solution for the response to a square wave may usually be substituted when the period required does not necessitate the computation of many terms of the Fourier series. The period $1/f_p$ chosen must be long enough to permit the transient response of the circuit under consideration to assume a substantially constant value between consecutive abrupt changes in the square wave.

If the square wave applied to the circuit is

$$E(t) = \frac{1}{2} + \frac{2}{\pi}(\sin 2\pi f_p t + \tfrac{1}{3}\sin 6\pi f_p t + \tfrac{1}{5}\sin 10\pi f_p t + \cdots + \frac{1}{n}\sin 2\pi n f_p t + \cdots \quad (1)$$

then the response is

$$e(t) = \frac{1}{2} + \frac{2}{\pi}(A_1 \sin 2\pi f_p(t - D_1) + \frac{A_3}{3}\sin 6\pi f_p(t - D_3) + \cdots + \frac{A_n}{n}\sin 2\pi n f_p(t - D_n) + \cdots \quad (2)$$

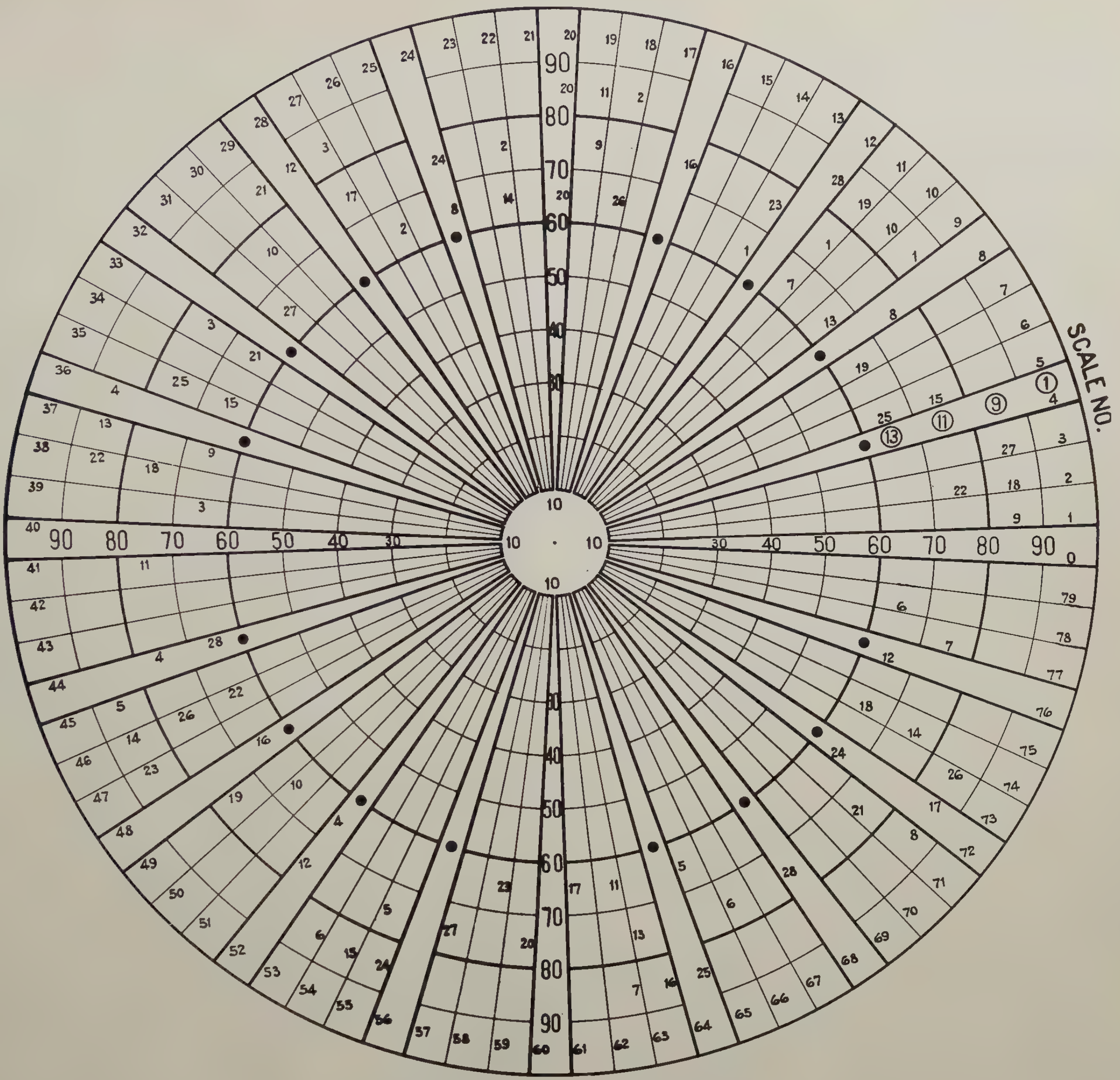


Fig. 12—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
1	0.25 Mc
9	2.25 Mc
11	2.75 Mc
13	3.75 Mc

where A_n and D_n are the amplitude response and delay of the circuit for the nth harmonic. D_n = phase shift in radians$/2\pi n f_p$.

Synthesis charts, Figs. 11, 12, and 13, were developed for the summing of the significant terms of the series (2) above. The principle of the charts is based on the vector representation of a sine function. For example, the term $A_n/n \sin 2\pi n f_p(t-D_n)$ is represented in Fig. 14(a) by the length of the perpendicular AC to the vector OD. The value of the term may be found rapidly, for specific values of t which have been determined in advance, by dividing the circumference of a circle into a number of equal parts.

For example, if the circle in Fig. 14(b) is divided into N equal segments, then the value of the sine term may be found every $1/Nnf_p$ second starting at $t=0$. It is convenient to designate a specific value of time t by one of the whole numbers between 0 and N.

A straightedge may be placed along the radial line which makes the required angle $2\pi n f_p D_n$ with the x axis. (See Fig. 14(b)). The quantity $2\pi n f_p D_n$ will usually be expressed in terms of the unit angular segment used in marking off the circumference of the circle. If a draftsman's triangle is moved along OD, successive values of the sine term corresponding to regular intervals of time may be read on a calibrated scale AC (marked on one leg of the triangle) by noting the points of intersection of the radial lines and an

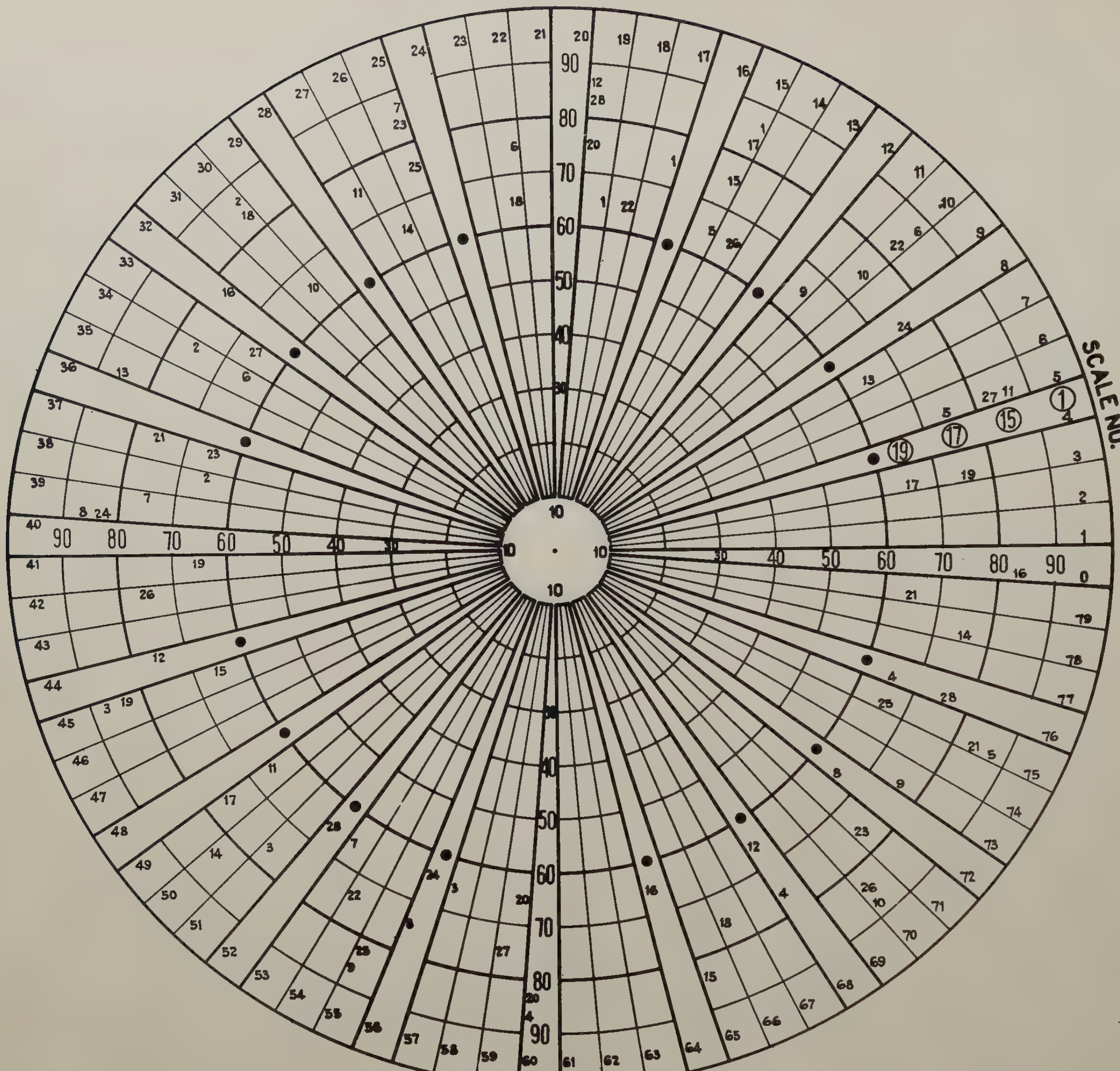


Fig. 13—Chart for square-wave synthesis.

Scale	0.05-microsecond intervals
(1)	0.25
(15)	3.75
(17)	4.25
(19)	4.75

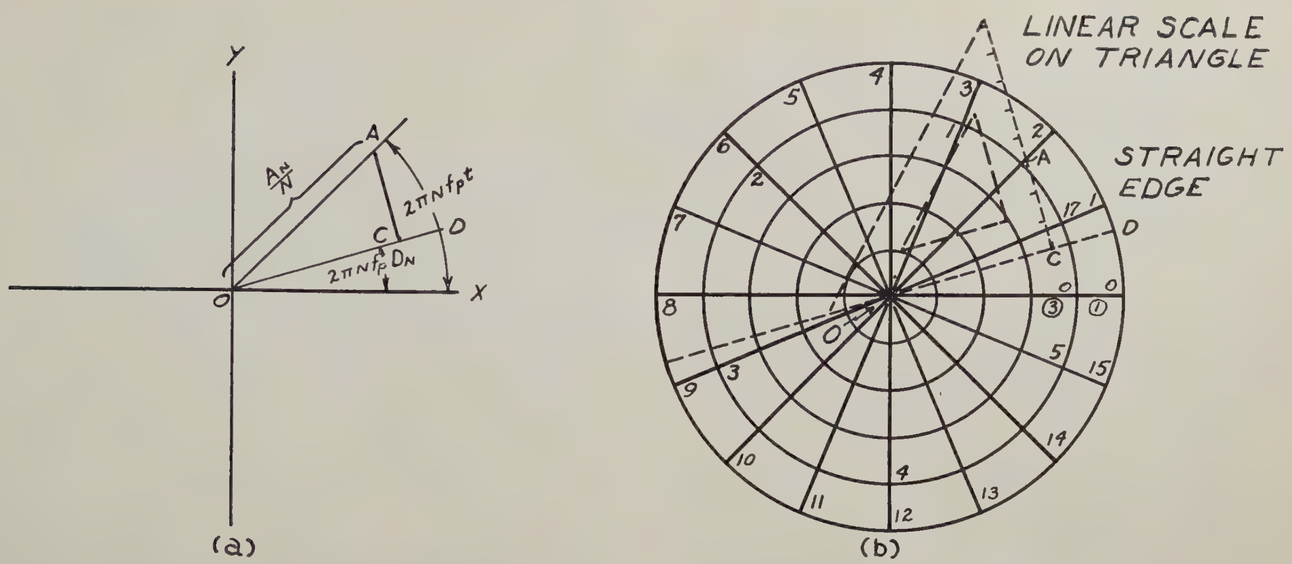
imaginary circle having a diameter OA equal to A_n/n. The multiplying factor $2/\pi$ in (2) may be taken into account by suitable calibration of the scale AC. A number of concentric circles permanently drawn on the chart, so as to divide the diameter of the largest circle into a convenient number of equal parts, will aid in finding the length OA as the triangle is shifted. Readings along the calibrated edge are positive above the line OD and negative below OD.

The same chart may be used for a number of different harmonic terms by simply numbering the radial lines appropriately for each term. Thus, in Fig. 14(b), if the circular scale (1) represents the fundamental term $\sin 2\pi f_p t$, then circular scale (3) will represent the 3rd-harmonic term $\sin 6\pi f_p t$ in which the unit angle is 3 times the unit angle in scale (1) etc. The steps involved in finding and summing the sine terms in (2) have been systematized in a set of synthesis charts, Figs. 11, 12, and 13. The following directions have been drawn up to facilitate rapid use of the synthesis charts.

A. Directions for Use of Synthesis Charts

Figs. 11, 12, and 13 are used in the compilation of the response of an electric circuit to a square wave as represented by the first 10 odd harmonics (1st, 3rd, . . . , 19th) when the sine-wave, phase-delay, and amplitude characteristics of the circuit are known.

The particular charts illustrated were arbitrarily



(a) Basic principle of construction of synthesis charts.

(b) Use of synthesis charts.

Fig. 14

designed for a fundamental frequency of 0.25 megacycle which has been found satisfactory for television purposes. In this instance the square-wave response is found at intervals of 0.05-microsecond (i.e., 1/80 of the period). The result is a close approximation to the response of a circuit to a unit function when (1) the square-wave response attains substantially a steady value within 2 microseconds (i.e., one half the period of the fundamental frequency) and (2) the contribution of harmonics beyond the 19th (4.75 megacycles) may be neglected.

The rapid use of the charts is expedited by observation of the following procedure in the preparing of Table II. A synthesis of the amplifier square-wave response shown in Fig. 8 from the theoretical amplitude and phase characteristics in Fig. 7 is used as an example.

TABLE II

Synthesis of 32-Stage Amplifier

(a) Order of Harmonic, n	(b) Frequency of Harmonic (Mc)	(c) Amplitude Response of Harmonic $\div n$ (in per cent)	(d) Delay of Harmonic $\times 20n$	Contributions of Harmonics Read from Charts				
				(0) $t=0$ microseconds	(1) $t=0.05$	(2) $t=0.1$	(27) $t=1.35$	(28) $t=1.4$
1	0.25	100	1.5	−7.5	−2.5	+2.5		
3	0.75	34	4.7	−7.8	−2.9	+2.3		
5	1.25	21	8.5	−8.4	−3.6	+1.6		
7	1.75	15	12.9	−8.1	−4.3	+0.8		
9	2.25	11	17.6	−6.6	−4.3	+0.2		
11	2.75	7.5	22.2	−4.7	−3.7	−0.1		
13	3.25	5	26.0	−3.0	−2.8	0		
15	3.75	3.5	29.0	−1.7	−2.0	+0.2		
17	4.25	2.5	31.6	−0.9	−1.4	+0.7		
19	4.75	2	33.4	−0.5	−0.9	+0.4		
Add +50				+50	+50	+50		
Instantaneous response				+ 1	+22	+59		

(1) Record in column (c) the amplitude response in per cent corresponding to each harmonic divided by the order of the harmonic.

(2) Shift the real delay curve up or down but parallel to itself to any new position for which the minimum delay in the essential frequency band is equal to, or near, zero. Shifting (i.e., subtracting a constant delay) is a convenience in manipulating the charts but is not a necessity.

(3) Record in column (d) the real delay (in microseconds) (or shifted delay, if performed) of each harmonic multiplied by 20 times the order of the harmonic.

(4) Proceed as follows to enter the contributions of the various harmonics in columns (0) to (28):

(A) Pass a straightedge through the center of Fig. 11 along the radial line corresponding to a number on the circular scale (1) equal to the delay of the fundamental frequency recorded in column (d). Circular scale (1) refers to the series of radial lines numbered 0 to 80 in ring (1). The straightedge remains fixed until steps (B) and (C) below have been completed.

(B) On the radial line number 0 on scale (1) find a point which is at a distance from the center of the chart equal to the number in column (c). The length of the perpendicular from this point to the straightedge is the contribution of the fundamental frequency at $t=0$. If the point referred to is *below* the axis, the contribution is entered in column (0) of the table with a negative sign. If above, the contribution is entered with a positive sign. The perpendicular is measured with a calibrated triangle illustrated in Fig. 15. One leg slides along the straightedge while the calibrated leg forms the perpendicular.

(C) The contribution of the fundamental at $t=0.05$ microsecond, to be entered in column (1) opposite 0.25 megacycle, is found by locating the point 1 on scale (1) and proceeding as directed for point 0. Point 2 corresponds to 2×0.05 microsecond. In a similar manner, the first row of entries in Table II are made.

(D) In order to find contributions of the 3rd harmonic, pass the straightedge through the number on scale (1) corresponding to the delay factor from column (d). Use scale numbers on scale (3) in order to locate the perpendicular and proceed with the measurement of its length as directed above for the fundamental frequency.

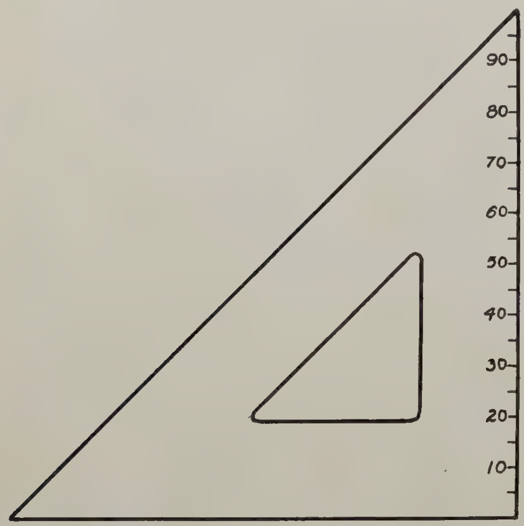


Fig. 15—Calibrated triangle[8] for reading contributions of harmonics from synthesis charts.

(E) Complete the table for the other harmonics in in a manner similar to the above.

(F) The instantaneous value of the square-wave response at $t=0$ microseconds is found by summing column (0) and adding 50 per cent. Similarly, the response is found at other times by summing the appropriate column and adding 50 per cent.

B. Examples of Synthesis of Square-Wave Response

Several points on the square-wave response of a 2-stage amplifier are synthesized in Fig. 8 from the amplitude and phase characteristics in Fig. 7. Since the theoretical response is shown, a direct indication of the accuracy of the synthesis is available. The discrepancy is seen to be only a few per cent in this example.

The chart method greatly simplifies and shortens the synthesis of square-wave response in complex cases in which rigorous mathematical formulas for response are impracticable. A typical example is the response of a 32-stage amplifier which was synthesized in Fig. 16 from the theoretical amplitude and phase characteristics of Fig. 17. In this particular case, an absolute measure of the error of the synthesis is not known. The deviation of any synthesized square-wave response from the exact square-wave response is due only to errors in manipulation of the charts and the neglect of contributions of harmonics beyond the limits of the charts (4.25 megacycles).

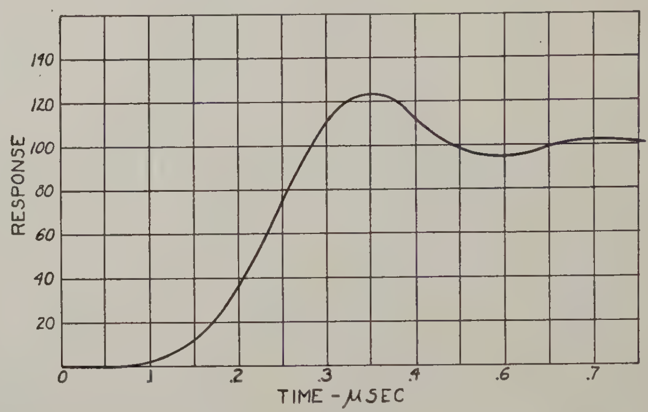


Fig. 16—Synthesis of the square-wave response of a 32-stage compensated resistance-capacitance amplifier in which $K=1.51$ and $f_0=7.7$ megacycles.

IV. Evaluation of the Square-Wave Response

Assume that a subject containing an extensive dark area and an extensive white area with a sharp vertical junction between the two is used as a test subject for a determination of the fidelity of transmission of a television system.[9] An ideal scanning device in crossing the junction would generate a unit voltage which becomes the input signal for a television transmission system under test. The output signal or response of the system may be symbolized in Fig. 18. The waveshape may be determined experimentally by a square-wave

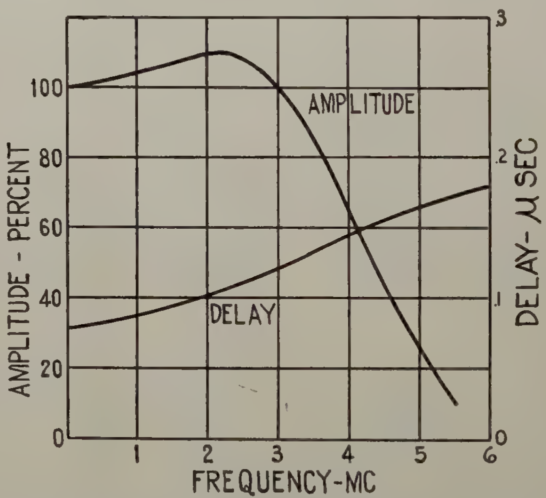


Fig. 17—Exact amplitude and delay characteristics of a 32-stage compensated resistance-capacitance amplifier in which $K=1.51$ and $f_0=7.7$ megacycles.

[8] The 100 per cent point on the radial scales of Figs. 11, 12, and 13 should coincide with the 63.6 $(=(\pi 2)100)$ per cent point on the calibrated triangle when the triangle is drawn to the same scale as the figures.

[9] The use of a single abrupt transition from one brightness to another brightness as a test subject for measuring "resolution" in a television picture was developed by R. D. Kell, A. V. Bedford, and G. L. Fredendall in "Determination of optimum number of lines in television system," *RCA Rev.*, vol. 15, pp. 7–30; July, 1940. A test pattern consisting of several converging bars is much more commonly used in evaluating an entire television system on account of convenience. The results obtained, however, are less significant because the resolution of the individual bars is not affected by the phase fidelity, whereas it is known that phase fidelity affects the sharpness and utility of most television pictures.

oscillograph or by synthesis from known sine-wave characteristics of the system. If the test subject were reproduced by an ideal scanning device (i.e., one having negligible aperture losses), actuated by the output signal, the variation of light intensity on the screen along the scanning lines would be as shown in the figure when using for the ordinates and abscissas the light intensity and distance, respectively. The distance required for the complete change from black to white is greater than zero due to the finite rate of rise of the response curve. Upon close observation of the screen, the junction would appear blurred. Furthermore, several alternate light and gray striations following the

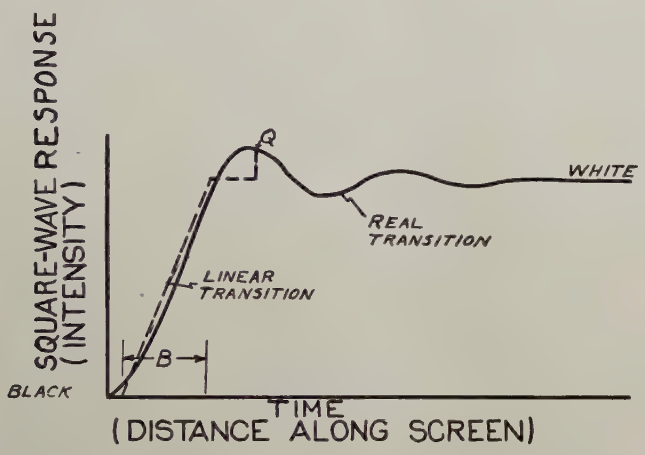


Fig. 18—Nonlinear response of a hypothetical complete television system and the equivalent linear response in terms of light along the screen surface or electrical response of the system.

junction would be observed as a consequence of the damped oscillation in the response. The overswing shown in Fig. 18 would not be entirely objectionable for television purposes because at the optimum viewing distance *most* of the overswing is not distinguishable from the transition and the net effect would be substantially a single transition. In fact, the visual sharpness of the transition is *enhanced* by the overswing in the response wave. If a quantitative measure of the sharpness were found, the response wave in Fig. 18, for example, could be compared directly with other waves having different shapes, and the relative merit of television apparatus with reference to picture sharpness could be determined.

With the object of finding such a measure of sharpness, the writers constructed several different synthetic black-white transitions with ink on cardboard using fine shading lines of variable widths to reproduce accurate half-tone values. When the transitions were viewed at a distance for which the "blur" was just discernible, that is, the optimum viewing distance it was observed that a simple *linear* transition similar to that shown dotted in Fig. 18 could be found for each nonlinear transition such that the observer was unable to distinguish between the nonlinear transition and its linear equivalent.

The width of the visually equivalent linear transition was termed the equivalent "blur" of the nonlinear transition. The blur may be specified in units of distance along the picture screen or in the corresponding units of time. Complex transition curves containing damped oscillations of sufficiently long duration are properly represented by a linear transition followed by that part of the oscillation which the eye does not include with the transition.

The comparison method of determining blur is objectionable because the labor involved is great and the evaluation of each transition depends somewhat upon the observer's judgment. A method not subject to these objections has been devised whereby the blur may be found directly from a plot of the light intensity along a transition. The steps involved in applying the method are given below in a "Generalized Statement" which defines the conditions under which a linear transition is visually equivalent to a nonlinear transition such as Fig. 18. For clarity, this "Generalized Statement" is presented in the form of a law or theorem, but we do not presume to use these terms until the statement has been proved more adequately by theory or experiment than is done in this paper.

Generalized Statement

A linear transition having a uniform rate of change of intensity along the surface from a first mean brightness to a second mean brightness is visually equivalent at the optimum viewing distance to any nonlinear transition

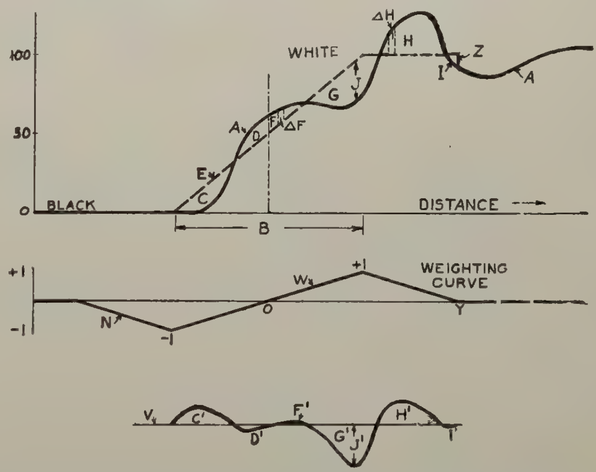


Fig. 19—Linear transition E is equivalent to A. By Condition 2, the algebraic sum of areas C, D, F, G, H, and I is zero. By condition 1, the sum of weighted areas C, D, F, etc., shown at C', D' F', etc., is zero

from the first mean brightness to the second mean brightness when conditions 1 and 2 below are fulfilled.

Condition 1—The summation of the weighted differences of the light intensities of the linear transition and of the nonlinear transition is zero, where the weighted differences are the real differences of light intensity along the transitions multiplied by a weighting factor. The weighting factor varies linearly with distance from a value of −1 at the first inflection point of the linear transition to +1 at the second inflection point; also linearly with

distance from a value of zero at a point preceding the first inflection point by half the distance between inflection points, to the −1 value at the first inflection point; also linearly with distance from the +1 value at the second inflection point to zero at a point following the second inflection point by half the distance between the inflection points; and is zero for all other points.

Condition 2—The summation of the differences of the light intensities along the transitions is zero, over the range where the weighting factor is not zero.

According to the generalized statement, the linear transition E in Fig. 19 is visually equivalent to the transition A. In this and other cases, it is convenient to consider the vertical dimensions J of the difference areas C, D, F, etc., as positive when A is above E and negative when A is below E. Then, if these dimensions are multiplied by the corresponding ordinates of the weighting curve N and plotted as at V, the positive and negative areas C', D', E', etc., are formed. Condition 1 is fulfilled if the algebraic sum of these areas C', D', F', etc., equals zero. Condition 2 simply requires that the algebraic sum of the positive and negative areas C, D, F, etc., equals zero.

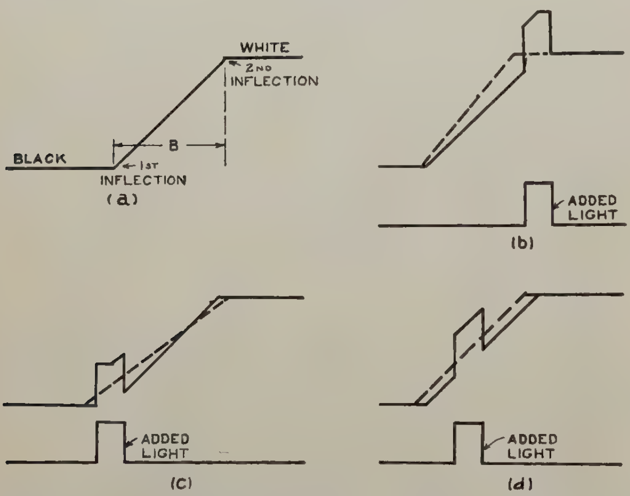


Fig. 20—Actual transitions are shown in solid lines and equivalent linear transitions are shown in dotted lines.

The location of a line E that fulfills the two conditions for a particular nonuniform transition is obtained by a trial-and-error method. The equivalent blur is indicated as the distance B. Each trial curve E requires a different weighting curve N since the weighting curve is itself defined by the equivalent linear-transition curve. Such a procedure would appear to be laborious since the line E has two variable characteristics, slope and position. However, the number of trials required is much reduced by the knowledge that for small changes of slope of line E only condition 1 is affected and for small changes of position only condition 2 is affected.

Although the generalized statement for defining a linear transition which is visually equivalent to a nonlinear transition is essentially empirical, the conditions appear to be consistent with observed properties of the eye. It is well known that the eye responds to a surface consisting of many tiny, uniformly distributed white dots on a black background (or black dots on a white background) as though the surface were uniformly white but the illumination reduced so that the total reflected light is the same. In other words, the eye responds to the average[10] brightness or to the total brightness. Condition 2 is consistent with this observation in that it requires equality of the total amount of light of the equivalent transition and the nonlinear transition.

The plausibility of condition 1 may be established by using the simple transitions of Fig. 20 to illustrate some basic factors which affect apparent steepness. Fig. 20(a) represents a linear change from black to white or from any half tone to another half tone. The blur by our definition is the distance B. Now assume that an extra amount of light is added near the second inflection point as shown in Fig. 20(b). As a result, the transition has been made effectively steeper so that the equivalent linear transition must also be made steeper as shown by the dotted line.

If the light were added near the first inflection point as shown in Fig. 20(c), it is apparent that the transition would be *less* steep as indicated by the dotted linear transition. From these two observations, it is plausible that the addition of light at the *middle* of the transition as in Fig. 20(d) would not alter the effective steepness. The equivalent linear transition, however, would be displaced to the left as required by condition 2.

If light were added at a point considerably ahead of the first inflection point of the linear transition or considerably after the second inflection point, it is reasonable that the effect on the steepness would be less than if the light were added near the inflection points.

Let us refer back to the example of Fig. 19 in which curve E is assumed to be the visual equivalent of the curve A. Obviously, E is *identical* to curve A except for the difference areas C, D, G, etc. Some of the areas such as C and H render curve A effectively steeper than the equivalent curve but other areas such as D and G detract from the steepness of curve A.

If curve E is to be the equivalent of curve A, the algebraic sum of the *effects* of all the aiding difference areas C, H, etc., must be canceled by the opposing areas. Condition 1 states this requirement and provides a weighting factor for determining the effectiveness of each difference area resulting from its location, as discussed in connection with Fig. 20.

A linear weighting curve has been taken arbitrarily due to the absence of evidence which would point to a specific form for the curve. A linear variation is simple and may also be considered as a mean between the various possible concave and convex forms.

It should be noted especially that in the application

[10] The eye also responds to the average brightness of a rapidly flickering source.

of the generalized statement to transitions which have appreciable irregularities in the region in which the weighting curve is zero (beyond Y), the irregularities are not included in the equivalent blur. As shown in Fig. 19, the equivalent curve may be continued through Z to the actual curve A. It should be noted also that the equivalent transition defined by the statement is intended to be equivalent only when viewed at a distance for which the nonlinear transition is indistinguishable from a linear equivalent transition. At such a distance, a very definite blur may be visible without the *shape* of the transition being discernible.

The observations made in the several paragraphs above lend support to the concept of an equivalent linear blur but obviously do not place the concept on a firm physical basis. Furthermore, the physical accuracy of the particular conditions set forth in the generalized statement have not been established because of inadequate knowledge of the eye and brain. However, actual viewing tests with specific transitions lead us to believe that the application of the generalized statement is sufficiently accurate to serve a useful purpose for the evaluation of the transitions found in television.

The authors feel that such a measure of blur also can have utility in specifying the quality of many other devices related to vision, such as lens system, photographic film and processes, facsimile transmission, printing, duplicating processes, and paper.

V. Application of Square-Wave Methods in Television

Square-wave and sine-wave measurements should be mutual aids in the solution of many television problems. In some instances, a square-wave measurement may furnish the data for analysis into amplitude and delay characteristics. In others, the evaluation of blur is indicated. In some applications, sine-wave measurement and synthesis of the square-wave response may be indicated, followed perhaps by evaluation. Most applications will not require the use of all three processes of square-wave treatment.

We know no simple satisfactory method of judging the degree of fidelity of a television system from inspection of the sine-wave characteristics. If the square-wave response of the system cannot be secured experimentally, resort must be had to synthesis from amplitude and phase data. However, acceptable tolerances in terms of sine-wave performance may be more easily determined when the amplitude and phase characteristics corresponding to a large variety of transient-response curves have been determined and cataloged.

In design work in which the characteristics of television apparatus must be determined by calculation based on circuit constants, synthesis from the sine-wave characteristics probably constitutes the only feasible means for obtaining the square-wave response. When the apparatus is susceptible to experimental test, the most expeditious method is direct measurement of the square-wave response by oscillographic equipment such as described in a companion paper.[3] An analysis for amplitude and phase characteristics may then be performed in order to facilitate the design of equalizing networks if required or for any other purpose. In this connection it is significant that the experimental difficulty of phase measurements of extensive apparatus such as a complete television system including transmitter and receiver by sine-wave methods is usually so great that the attempt is not often made.[11,12] Notwithstanding, a reasonably linear phase shift is conceded to rank with amplitude uniformity in importance. The extreme ease with which square-wave response can be recorded with suitable equipment has been pointed out in the companion paper. The analysis of the response for delay versus frequency through the use of charts is simple and immediate.

In particular cases, square-wave measurements may provide useful data which are almost impossible to obtain by sine-wave measurements and synthesis. An example is the modulation amplifier of a television transmitter in which the output impedance may be relatively high at low frequencies, in order to permit a high output voltage, and relatively low at high frequencies such that the frequency fidelity of the stage alone is poor. Adequate equalization may be inserted in earlier stages of the system but high-frequency components may be saturated at high levels. A sine-wave characteristic of the entire amplifier taken at a low level for which saturation is negligible would indicate high fidelity. It would be almost impossible to synthesize from the low-level sine-wave data the transient response for a high level corresponding to conditions occurring in common use. Square-wave oscillographic tests, however, would indicate the transient response corresponding to any desired level. An evaluation of the square-wave response for the purpose of finding the blur corresponding to various levels would be significant but a determination of the sine-wave response at levels where saturation exists would have no meaning in the usual sense.

In general, square-wave methods have greatest usefulness (as compared with sine-wave methods) in dealing with performance of units which are likely to contribute a substantial amount of the distortion in the transmission characteristic of the system. Included in this classification are entire transmitters, entire receivers, long transmission lines, pickup chains, and any single amplifier stages which may be regarded as "bottleneck" stages. It would be rather pointless to find the square-wave response and the equivalent blur of a single stage of a video voltage amplifier of good

[11] B. D. Loughlin, "A phase curve tracer for television," Proc. I.R.E., vol. 29, pp. 107–115; March, 1941. Loughlin describes apparatus which furnishes a direct plot of phase versus frequency on the screen of a cathode-ray tube. The complexity of the apparatus may limit its general utility.

[12] M. E. Strieby and J. F. Wentz, "Television transmission over wire lines," *Bell Sys. Tech. Jour.*, vol. 20, pp. 62–81; January, 1941.

fidelity in a system in which many similar stages exist. Usually the distortion of a single stage is so small that only the accumulated effect of several stages is clearly evident in the over-all square-wave response. The writers are not aware of the existence of a practicable method of combining the individual square-wave response of two or more units for the purpose of finding the over-all square-wave response. Even if a feasible method should exist, the experimental errors involved in many individual measurements would jeopardize the dependability of the calculated over-all response.

If units of good fidelity must be considered individually, it becomes expedient to determine first the sine-wave characteristics and then to combine the sine-wave characteristics so that a synthesis of the over-all square-wave response may be performed.

The accuracy of square-wave methods is more than adequate to reveal those imperfections of transmission which cause discernible effects in a television picture. Therefore, the authors hope that this paper will be a contribution to the general problem of working out electrical specifications for television transmitters and other television apparatus, giving information regarding the steepness of rise and the amplitude of overswing of the square-wave response.

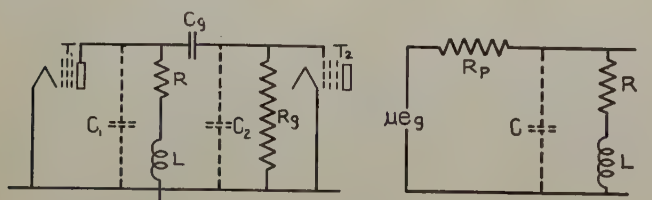
Appendix I

Analysis of Square-Wave Response

Let the square-wave response be approximated by a stepped wave f as shown in Fig. 1. The kth step may be approximated by the series

$$A_k\left[\frac{1}{2}+\frac{2}{\pi}\sum_{1}^{N}\frac{1}{n}\sin n\omega_o(t-T_K)\right] \tag{3}$$

where N is very large.



(a) Compensated resistance-capacitance amplifier. (b) Equivalent circuit for high frequencies.

Fig. 21

The sum of M steps leads to the following expression for the stepped wave

$$\sum_{m=1}^{M}\frac{A_m}{2}+\frac{2}{\pi}\sum_{m=1}^{M}\left[A_m\sum_{n=1}^{N}\frac{1}{n}\sin n\omega_o(t-T_m)\right]. \tag{4}$$

The constant term in (3) is not of interest. If the order of summation of the second term is reversed, there results

$$\frac{2}{\pi}\sum_{n=1}^{N}\frac{1}{n}\sum_{m=1}^{M}\left[A_m\sin n\omega_o(t-T_m)\right]. \tag{5}$$

The inner sum written for some specific value of n as N_1 is the total N_1th harmonic content of the stepped wave. That is,

$$\begin{aligned} E_{N_1} &= \frac{2}{n}\frac{1}{N_1}\sum_{m=1}^{M}A_m\sin N_1\omega_o(t-T_m) \\ &= \frac{2}{\pi}\frac{1}{N_1}B_{N_1}\sin N_1\omega_o(t-t_{N_1}). \end{aligned} \tag{6}$$

Since the input square-wave signal may be approximated by the form

$$\frac{1}{2}+\frac{2}{\pi}\sum_{n=1}^{N}\frac{1}{n}\sin n\omega_o t$$

it follows that the N_1 component of the input is

$$\frac{2}{\pi}\frac{1}{N_1}\sin N_1\omega_o t.$$

The amplitude response of the circuit is, therefore, B_{N_1} and the time delay is T_{N_1}. These two quantities may be found graphically as indicated in Fig. 1(a) and (b).

Since ω_o may be allowed to approach zero, it follows that reference to a fundamental frequency is not required and the amplitude and phase corresponding to any frequency may be found.

Appendix II

The Compensated Resistance-Capacitance Amplifier

A schematic diagram of the amplifier appears in Fig. 21. When $R_p \gg Z(L, C, R)$ the response of the equivalent circuit to a unit function is expressed by the following equation:

$$e=\frac{\mu R}{R_p}\left[1-e^{-\pi f_o K t}\{A\sin(\Omega_1 t+\psi)+Bt\sin(\Omega_1 t+\beta)\}\right]$$

in which

$$t = \text{time} \qquad \Omega_1 = 2\pi f_o\sqrt{1-K^2/4}$$

$$f_o = 1/(2\pi\sqrt{LC}) \qquad B = \sqrt{P^2+N^2}$$

$$A = \sqrt{1+M^2} \qquad \psi = \tan^{-1}\frac{1}{M}$$

$$K = 2\pi f_o RC$$

$$M = \frac{-3+3K^2-K^4/2}{4K(1-K^2/4)^{3/2}} \qquad \beta = \tan^{-1}\frac{N}{P}$$

$$N = \frac{\pi f_o(K^2-1)}{K^2\sqrt{1-K^2/4}} \qquad P = \frac{\pi f_o(3-K^2)}{2K(1-K^2/4)}.$$

The amplitude response is

$$\text{amplitude} = \frac{\mu R}{R_p}\sqrt{\frac{1+B^2/K^2}{B^2K^2+(B^2-1)^2}}$$

$$\text{delay} = \frac{1}{2\pi f}\tan^{-1}(1-K^2-B^2)B/K$$

in which

$$f = \text{frequency}$$
$$B = f/f_0$$
$$K = 2\pi f_0 RC.$$

Appendix III

Relation Between Square-Wave Analysis and Fourier Analysis

Assume that the response of a linear electrical network to a unit function is represented by $e_1(t)$ in Fig. 22(a). The response $e_2(t)$ of the same circuit to a square pulse Δt seconds long is found by shifting $e_1(t)$ to the right along the time axis by an amount Δt and subtracting the result from $e_1(t)$ that is, $e_2(t) = e_1(t) - e_1(t+\Delta t)$. If the condition is imposed that $e_1(t) = 1$ for $t \geqq t_q$, then $e_2(t) = 0$ for $t \geqq (t_q + \Delta t)$.

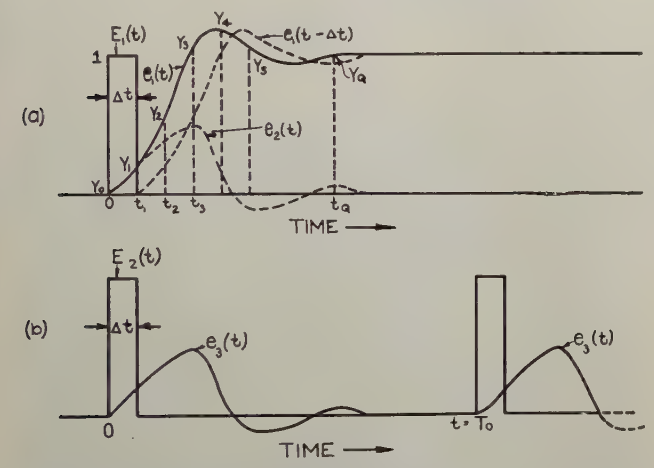


Fig. 22—Fourier analysis applied to the response of a circuit to a square pulse compared with square-wave analysis applied to the response to a unit function.

It is clear that the waveform of the response $e_3(t)$ of the same circuit to the periodic pulse wave of Fig. 22(b) is given for each cycle by $e_2(t)$ if $T_0 \geqq (t_q + \Delta t)$. Hence, the coefficients of the Fourier series written for $e_3(t)$ may be expressed in terms of the ordinate readings of $e_1(t)$ taken at successive equal time intervals. Thus

$$\begin{aligned} e_3(t) = a_0 &+ a_1 \cos 2\pi f_0 t + \cdots + a_k \cos K2\pi f_0 t + \cdots \\ &+ a_{n/2} \cos \frac{n}{2} 2\pi f_0 t + b_1 \sin 2\pi f_0 t + \cdots \\ &+ b_k \sin K2\pi f_0 t + \cdots \\ &+ b_{(n/2)-1} \sin \left(\frac{n}{2} - 1\right) 2\pi f_0 t \end{aligned}$$

where

$$\begin{aligned} a_k = 2\Delta t f_0 [&(y_1 - y_0) \cos K2\pi\Delta t f_0 \\ &+ (y_2 - y_1) \cos K4\pi\Delta t f_0 + \cdots \\ &+ (y_q - y_{q-1}) \cos Kq2\pi\Delta t f_0] \\ = 2\Delta t f_0 A& \end{aligned}$$

and

$$\begin{aligned} b_k = 2\Delta t f_0 [&(y_1 - y_0) \sin K2\pi\Delta t f_0 \\ &+ (y_2 - y_1) \sin K4\pi\Delta t f_0 + \cdots \\ &+ (y_q - y_{q-1}) \sin k_q 2\pi\Delta t f_0] \\ = 2\Delta t f_0 B.& \end{aligned}$$

The Fourier series for $E_2(t)$ is

$$E_2(t) = \frac{T_0}{\Delta t} + \frac{2}{\pi} \sum_K \frac{1}{k} \sin K\pi \frac{\Delta t}{T} \cos 2\pi K f_0 t.$$

Hence, the amplitude characteristic of the circuit is given by

$$R(K, f_0) = \frac{\pi \Delta t K f_0}{\sin \pi \Delta t K f} \sqrt{A^2 + B^2}$$

$$\text{or } \lim_{f_0 \to 0} R(K, f_0) = R(f) = \frac{\pi \Delta t f}{\sin \pi \Delta t f} \left[\lim_{f_0 \to 0} \sqrt{A^2 + B^2} \right]$$

and the delay characteristic is given by

$$\lim_{f_0 \to 0} \frac{\theta(K, f_0)}{2\pi f} = \frac{1}{2\pi f} \tan^{-1} \frac{B}{A}.$$

Now $\sqrt{A^2+B^2}$ is the same expression which is obtained for the amplitude characteristic by the graphical method of square-wave analysis. Hence, $\pi\Delta t f/\sin \pi\Delta t f$ may be regarded as the conversion factor which converts square-wave analysis into Fourier analysis when the latter is based on the square pulse.

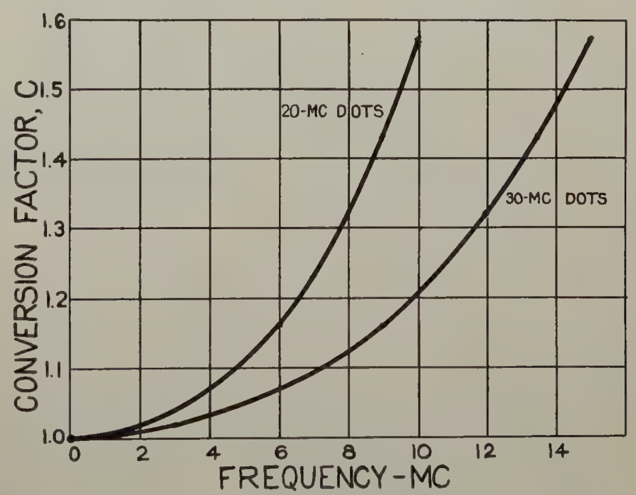


Fig. 23—Conversion factor for obtaining Fourier analysis from square-wave analysis. C times amplitude according to square-wave analysis equals amplitude according to Fourier analysis.

The magnitude of the conversion factor is shown in Fig. 23. It may be concluded that for most applications of square-wave analysis in television, the conversion factor may be taken equal to 1. The delay characteristics as determined by square-wave analysis and Fourier analysis are identical.

A Portable High-Frequency Square-Wave Oscillograph for Television*

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Summary—*A portable high-frequency oscillograph for television is described by which a square-wave (100-kilocycle) response may be viewed as a dotted wave and readily recorded as a series of readings. The dots are spaced at 1/30- (or 1/20-) microsecond intervals. No electrical connection is required between the oscillograph and the square-wave generator other than that established through the apparatus under test since the synchronous sweep and timing dots are derived from the square-wave response of the apparatus. Circuit diagrams of the square-wave generator and square-wave oscillograph are given.*

I. Introduction

The applications of square-wave measurements in television are developed in detail in a companion paper[1] in which the following aspects are treated: (1) a graphical chart method for analyzing the response of a system to a square-wave input signal to obtain the sine-wave amplitude and delay characteristics; (2) a graphical chart method for synthesizing the response to a square-wave from the sine-wave amplitude and delay characteristics; (3) a method for evaluating the mean steepness of a transient-response wave in terms of the width of blur produced in a television image by a visually equivalent wave having a linear change between two reference half tones; and (4) specific applications of square-wave measurements. When an experimentally determined square-wave response is required for the execution of items (1), (3), and (4), the use of a suitable oscillograph becomes necessary. It is the purpose of the present paper to describe a portable high-frequency oscillograph and a square-wave generator by which a square-wave response may be viewed as a dotted wave and readily recorded as a series of amplitude readings separated by known equal time intervals.

II. Choice of the Fundamental Frequency of the Square Wave

It can be shown that if the response of an electrical circuit to a unit-function input subsides to a substantially constant value after the transient interval as shown in Fig. 1 (A), the response of the same circuit to a square wave as shown in Fig. 1 (B) contains a transient interval or transition $a-b$ essentially similar to that due to the unit function. In practice, the transitions may be considered identical when the period of the square wave is about 3 or more times greater than the duration of the transient interval as in Fig. 1. For most television measurements, a 100-kilocycle square wave has been found to be adequate. Hence, the design of a square-wave oscillograph was based on this fundamental frequency. The analysis and evaluation processes mentioned in Section I presuppose that the square-wave response of the apparatus under test is

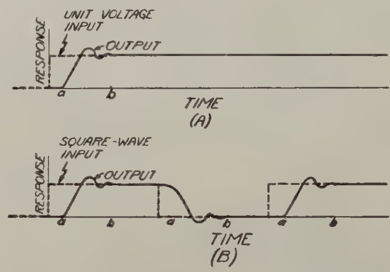


Fig. 1—(A) Response of a circuit to a unit function.
(B) Response of the same circuit to a square wave.

essentially equivalent to the unit-function response so far as the high-frequency components are concerned.[2]

III. The Essential Components of the Oscillograph

Fig. 2 is a block diagram of a typical test arrangement including the 100-kilocycle square-wave generator, the apparatus under test, and the square-wave oscillograph. The essential components of the oscillograph are shown as blocks within a larger block. No electrical connection is required between the oscillograph and the generator other than that established through the apparatus under test because the synchronous sweep and timing signals are derived from the square-wave output of the apparatus.[3] This feature contributes to convenience even in a very simple setup, but is of greatest advantage in the testing of distributed systems in which the input and the output are separated by a considerable distance as in a complete transmitter and an associated television receiver.

Fig. 3 is a photograph of the square-wave oscillograph and square-wave generator. In Fig. 4, the probe

* Decimal classification: R388×R583. Original manuscript received by the Institute, February 19, 1942. Presented, Summer Convention, Detroit, Michigan, June 25, 1941.

† RCA Manufacturing Company, Camden, New Jersey.

[1] A. V. Bedford and G. L. Fredendall, "Analysis, synthesis, and evaluation of the transient response of television apparatus," Proc. I.R.E., this issue, pp. 440–457.

[2] The use of a 60-cycle square-wave generator and an oscillograph for investigation of the behavior of a television system at *low frequencies* of the order of the field-scanning rate is well established. In these measurements, performance is judged by inspection of the "tilted" output wave and harmonic analysis of the wave is usually not desired. In general, a television system will have uniform response over a frequency range of many octaves in the region between the so-called "low-frequency" end and the "high-frequency" end, so that fidelity measurements at the two ends of the spectrum may be considered separately. This paper will be concerned only with the high-frequency end.

[3] Most previous square-wave oscillographs have depended upon a linear sweep and photography for recording. Some have required additional connections between the generator and the cathode-ray tube for timing. Such an oscillograph is described by H. E. Kallmann in "Portable equipment for observing transient response of television apparatus," Proc. I.R.E., vol. 28, pp. 351–360; August, 1940.

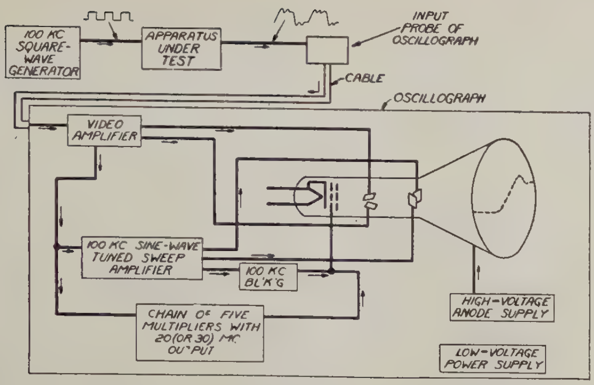


Fig. 2—Block diagram of a typical test arrangement showing the square-wave generator, apparatus under test, and the square-wave oscillograph. Internal components of the oscillograph also appear in block form.

and power cord have been removed from the storage compartment in front of the cathode-ray tube. The theory of operation of the various components together with certain constructional details is given below.

Fig. 3—Square-wave oscillograph and square-wave generator.

A. Input Probe and Video Amplifier

Fig. 5 shows a schematic diagram of the probe and the video amplifier for vertical deflection of the cathode-ray tube. Rectangles enclose all external controls on the diagram. The input probe contains an

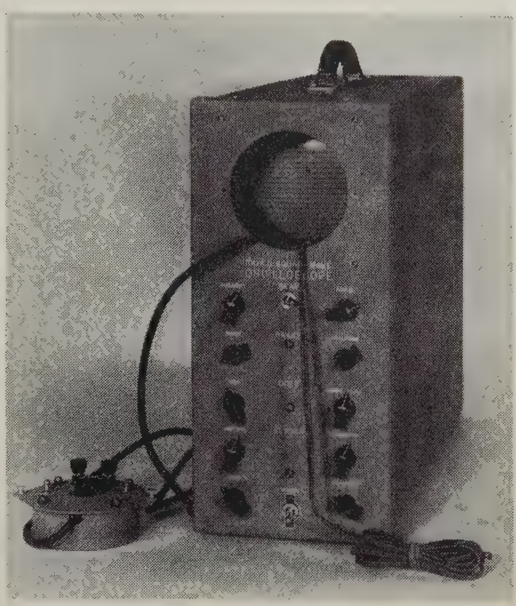


Fig. 4—Square-wave oscillograph and probe.

attenuator and a control for varying the internal capacitive load across the input terminals. In order to allow convenient use with short leads from the apparatus under test, the probe is in a small metal box connected to the oscillograph proper by a permanently shielded cable.

In experimental work, a record of the square-wave response at the grid of a tube or at the grid of a kinescope where the impedance is high is frequently desired. For such applications, the probe may be substituted for the input capacitance of the tube or kinescope and the total capacitance adjusted to the original value. A satisfactory substitution can generally be made since the minimum input capacitance of the probe is only 8 micromicrofarads. The resistive component of input impedance is negligible.

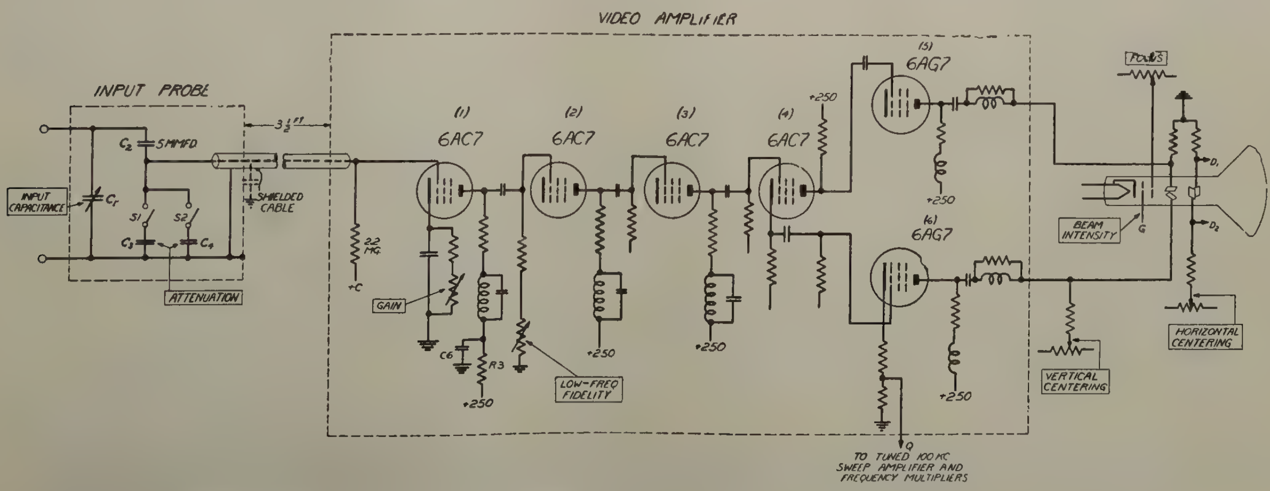
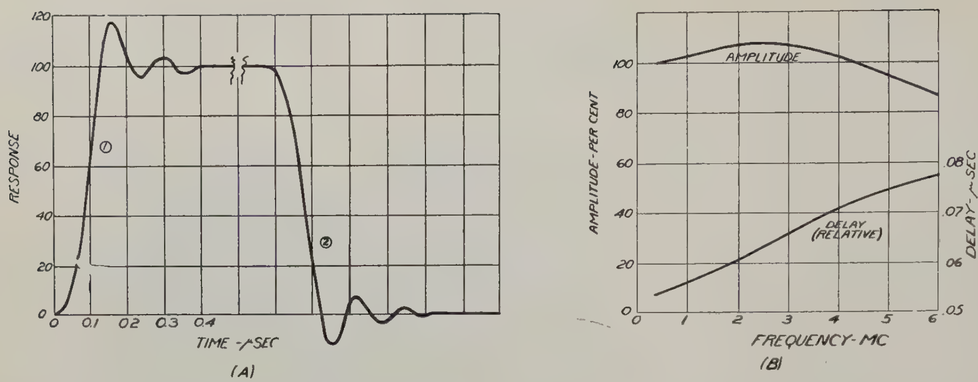


Fig. 5—Simplified circuit diagram of probe, video amplifier, and cathode-ray tube in the square-wave oscillograph.



(A) Square-wave response of generator and oscillograph only.

(B) Combined amplitude and delay characteristics of generator and oscillograph according to chart analysis of rising trace in (A).

Fig. 6

When switches S_1 and S_2 are open, the capacitance C_2 and the distributed capacitance of the probe cable comprise a capacitive potentiometer. The attenuation present when S_1 and S_2 are open was not purposely inserted but is incidental to the primary object of a low input capacitance. When the signal to be measured has a high voltage level, the capacitor C_3 or capacitor C_4 may be switched across the cable to attenuate the signal by a factor of 3 or 9 times as desired. The effect of the high-resistance grid leak of the first amplifier stage upon frequency fidelity of the capacitive potentiometer is negligible for frequencies above 250 kilocycles (lower limit used in the analysis charts). The grid-leak resistance, which is higher than commonly satisfactory for use with the 6AC7 type of tube, is low enough for stable operation in this application because the fixed portion of the cathode self-bias resistor is much higher than normal. The *positive* bias voltage applied to the lower end of the grid leak provides a net bias on the tube that is essentially normal, but does not interfere with the desirable automatic-control voltage generated across the high cathode resistor. The variable portion of the cathode resistor provides additional bias for fine control of the mutual conductance of the tube.

The plate circuits of the first three video stages are simple inductance-compensated stages. The fourth stage delivers cathode-output and plate-output signals for driving the push-pull output tubes. This stage requires no high-frequency equalizing circuits and is also free from amplitude distortion as a consequence of the large amount of degeneration.

Sufficient gain or attenuation control is available so that a peak-to-peak deflection of 100 units on the oscilloscope screen is possible throughout an input range of 1 to 100 volts, peak-to-peak. This range is adequate for most square-wave measurements on video apparatus in the television receiver and transmitter without the use of auxiliary amplifiers or attenuators.

The frequency response of the probe and amplifier is essentially flat from 100 kilocycles to 6 megacycles. A phase characteristic of the probe and amplifier is difficult to obtain independently of the phase characteristics of the square-wave generator. Fortunately, the really pertinent data are not the characteristics of the component parts but the over-all amplitude and delay characteristics of the square-wave generator and the oscillograph. Fig. 6 shows these characteristics as determined from a chart analysis of the square-wave response of the video amplifier and square-wave generator when directly connected. The amplitude characteristic is flat within ± 4 per cent out to 4.5 megacycles and the corresponding delay characteristic is flat within ± 0.01 microsecond. If interest is restricted to the usual video spectrum of 4.5 megacycles, these characteristics allow use of the equipment without corrections where only moderate accuracy is required. When corrections as indicated in Fig. 6 (B) are applied, the frequency range and the accuracy may be considerably extended.

B. Sweep Circuit for Horizontal Deflection

The sweep circuit is shown in Fig. 7. Tube 8 is driven by the 100-kilocycle-wave signal which exists across a part of the cathode resistor (point Q, Fig. 5)

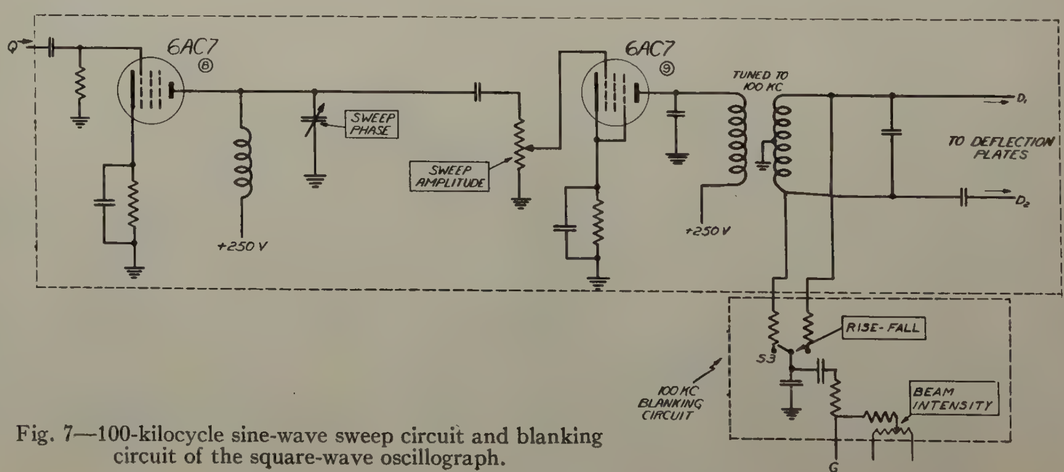


Fig. 7—100-kilocycle sine-wave sweep circuit and blanking circuit of the square-wave oscillograph.

of one output tube of the video amplifier. When the gain of the amplifier has been adjusted for a suitable vertical deflection on the cathode-ray-tube screen, the square-wave signal at point Q is also suitable for driving the sweep circuit.

The plate circuits of tubes (8) and (9) are tuned to 100 kilocycles; hence, the horizontal-deflection voltage appearing across D_1-D_2 is essentially a sine wave. The voltage across D_1-D_2 is used to deflect the cathode-ray tube horizontally as indicated by the similarly marked points in Fig. 5. The "sweep amplitude" is usually adjusted to several times the value which would just traverse the screen so that the visible portion of the sweep is reasonably linear. There is no need for the usual saw-tooth sweep circuit and synchronizing circuit since the use of timing dots, as explained below, obviates the necessity of a *linear* time axis. Centering of the square-wave response on the cathode-ray-tube screen is accomplished by (1) adjustment of the tuning condenser or "sweep-phase" control in the plate circuit of tube (8) and (2) by adjustment of the "horizontal-centering" control in the cathode-ray-tube circuit. The former control alters the phase relation between the sine-wave sweep signal and the square-wave signal across the vertical deflection plates. This control covers a large range and serves a very useful purpose as explained below.

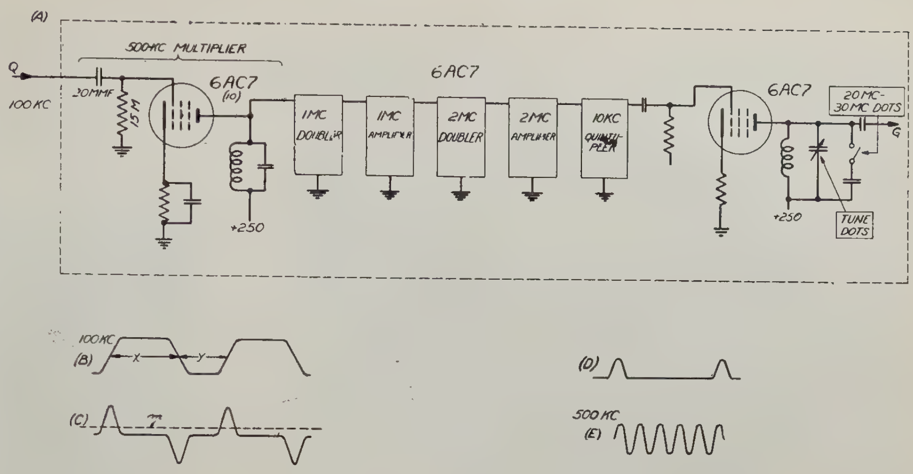


Fig. 8—(A) Dotter circuit of square-wave oscillograph.
(B) Square-wave signal at point Q.
(C) Differentiated signal at grid of 500-kilocycle multiplier tube.
(D) Plate current of multiplier tube.
(E) Output voltage of 500-kilocycle multiplier.

C. Cathode-Ray Keying Circuit for Timing

Almost every significant interpretation of a square-wave response curve, such as the evaluation of picture detail transmitted or the determination of amplitude and delay characteristics, depends upon the time base of the response curve as well as upon its shape. Therefore, an accurate means for measuring time along the curve appearing on the fluorescent screen of the cathode-ray tube is indispensable. The required accuracy is accomplished by a keying system which extinguishes the cathode-ray beam at either a 20- or a 30-megacycle rate as desired. The keying or "dotter" circuit shown in Fig. 8 (A) receives the amplified 100-kilocycle square wave from point Q in Fig. 5. The high-frequency keying signal which keys the beam is then generated by means of the chain of 5 frequency multipliers. Since the keying signal is synchronous with the square wave and the sweep circuit, the wave traced upon the screen consists of a dotted line in which the time interval between dots is the known time for 1 cycle of the keying signal. In order that the plate circuit of the first multiplier tube (10) shall have an abundant amount of 5th-harmonic signal, independent of the ratio X/Y in the square wave shown in Fig. 8 (B), the square-wave signal from Q is effectively differentiated by the coupling condenser and grid leak and then clipped by the tube (10) as shown in Fig. 8 (C) and (D).

In an earlier arrangement, the 100-kilocycle sine-wave sweep signal was impressed on the chain of multipliers. This method was not satisfactory because small spurious changes in the video input level and in the amplification caused the amount of saturation of the first multiplier to vary, thereby changing the relative phase position of the harmonics with respect to the transition point in the square-wave test signal. Although the phase changes represented very small time changes compared with the period of the 100-kilocycle wave, the effect was great enough to cause the dots on the oscilloscope trace to move along the wave over distances comparable to the space between dots.[4] Since the reading and recording of the ordinates of the dots comprising the trace on the screen requires appreciable time, spurious changes in the phase of the dots seriously impaired the accuracy of recording. The improved method of deriving the dots described above provides very stable timing because the pulses of plate current shown above line m in Fig. 8 (C) are fixed with respect to the rising portion of the square wave in Fig. 8 (B).

In each unit of the chain of conventional frequency multipliers, the amplitude of the input sine wave is sufficient to cut off the tube. The proper harmonic is then selected by a parallel-tuned plate circuit. The plate circuit of the last multiplier may be tuned to either 20 or 30 megacycles depending upon the position

[4] This phenomenon, by which a small change in phase in an early stage of a chain of frequency multipliers causes a large phase change in the final high-frequency stage, is used to advantage in the Armstrong method of frequency modulation for sound transmission.

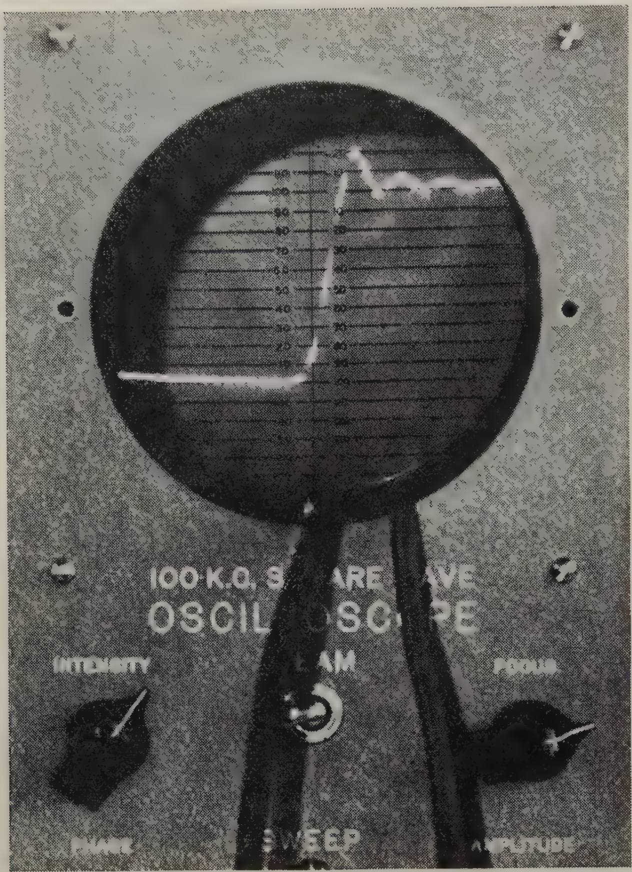


Fig. 9—Actual view of a square-wave response showing rising trace.

of the switch marked "20-Mc—30-Mc Dots." The variable condenser marked "Tune Dots" serves as a vernier. The output point G of the dotter circuit is connected to the grid of the cathode-ray-tube as shown in Fig. 5. Small and well-defined dots are obtained when the bias of the cathode-ray-tube is sufficiently negative to extinguish the beam except during positive peaks of the keying signal.

D. 100-Kilocycle Blanking Circuit

In order to allow easy observation of either the rising or the falling trace of the square-wave response (one at a time), a blanking circuit is provided, as shown in Fig. 7, for extinguishing the beam during the undesired trace. One trace is extinguished when the grid of the cathode-ray tube is biased by a blanking signal derived from the voltage at *one* horizontal-deflection plate and shifted nearly 90 degrees by a resistance-capacitance network. The other trace is extinguished by throwing switch S_3 which applies a blanking signal derived from the voltage at the other deflecting plate and shifted by a similar resistance-capacitance network. Both traces may be viewed at the same time if the "beam-intensity" control is raised above the normal level. Fig. 9 shows only the rising trace. The output circuit at G, in Fig. 8 which connects the grid of the cathode-ray-tube, has impedance high enough not to interfere with the high-frequency keying of the grid.

E. Mechanical Features

The chassis of the oscillograph extends vertically along the length of the case and is fastened to the center of the front control panel as shown in Fig. 10.

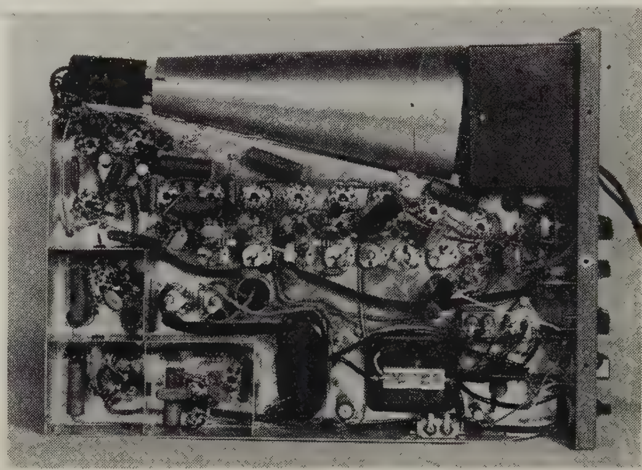


Fig. 10—Chassis of the square-wave oscillograph.

All electrical parts are mounted so that the instrument may be operated when removed from the case for servicing. All tubes are mounted on one side of the chassis and all small parts and wiring are accessible from the other side. The cathode-ray-tube is mounted with the face recessed several inches behind the viewing aperture in the control panel, thus forming an effective light shield.

Provision is made for storing the probe and the power cord in the compartment in front of the cathode-ray tube. The oscillograph weighs only 46 pounds and is readily portable.

IV. The Square-Wave Generator

An ideal square wave with infinite steepness contains all odd harmonics up to infinite frequency. The amplitudes of the harmonics are inversely proportional to the frequency and the phases are such that all harmonic waves pass through zero at the same time as the fundamental. High-frequency cutoff, reduction of the high-frequency components, or relative phase shift tends to reduce the steepness of rise and fall of the square wave.

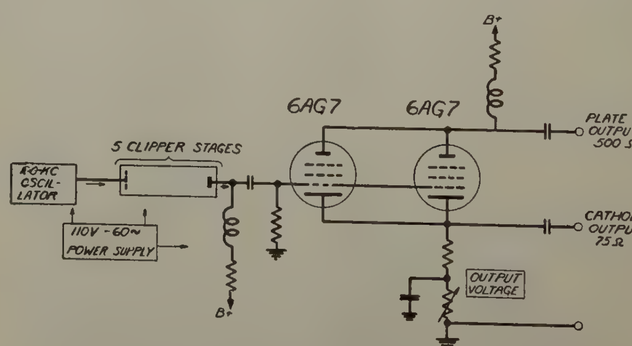


Fig. 11—Block diagram of square-wave generator.

The 100-kilocycle square-wave generator shown in the block diagram in Fig. 11 employs a sine-wave oscillator followed by several stages of limiting amplifiers which limit as a consequence of plate-current cutoff. If distributed circuit capacitance were not a controlling factor, a square wave of any prescribed steepness could be derived by repeated limiting and amplifying of a sine wave.

Since the square-wave generator for testing with the square-wave oscillograph must have good fidelity up to at least 4.5 megacycles (preferably higher), it was found necessary to provide inductance compensation for the interstage coupling between the later limiter stages. The generator has a 75-ohm output circuit supplied from the cathode of the last stage and a 500-ohm high-level output circuit supplied from the plate of the same stage. The voltage output is controlled by varying the bias on the output stage. The weight of the generator is 16 pounds.

V. Operation of Square-Wave Measuring Equipment

The arrangement of the square-wave measuring equipment for test is shown in Fig. 2. A choice between cathode-output and plate-output terminals of the square-wave generator is dictated by the nature of the input impedance of the apparatus under test. In instances which permit a 500-ohm source, the plate output may be used with a greater available maximum amplitude.

The probe is connected at the point in the apparatus at which the square-wave response is required. If this point has high impedance, it is desirable to disconnect some circuit element such as a tube grid and then adjust the probe capacitance to give the original total capacitance at the point as indicated by a capacitance meter. If no circuit element can be removed, the probe capacitance should be made a minimum.

Three controls may be manipulated in order to secure a full-scale square-wave response trace on the cathode-ray tube: (1) output control of the square-wave generator, (2) gain control of the oscillograph, and (3) attenuator switches in the probe. The manner in which a full-scale deflection is obtained is dependent on the input and output voltage conditions desired for the apparatus under test. Horizontal and vertical centering controls permit centering of the trace on the calibrated scale of the cathode-ray tube. For convenience in recording data, the two constant levels of the trace, preceding and following the abrupt transition, should lie on the 0 and 100 per cent lines of the calibrated scale as shown in Fig. 12.

The choice of 20- or 30-megacycle dots is dependent upon the use to be made of the data. The use of 30-megacycle dots defines the response wave more completely so far as high-frequency components are concerned; thus allowing greater accuracy in the analysis of the response for determination of the sine-wave characteristic of the apparatus under test. However, the use of 20-megacycle dots is desirable for quick recording and analysis where less accuracy is acceptable.

A record of the response consists merely of a sequence of numbers or ordinates corresponding to the vertical positions of the dots on the trace. The first reading is 0 when the vertical centering is properly adjusted and corresponds to the dot which just precedes the beginning of the rise (or fall) of the trace. In order to obtain the second reading, the entire trace is shifted by means of the "sweep-phase" control until the next or second dot occupies a position on the calibrated vertical axis as shown in Fig. 12 (A). The corresponding ordinate is recorded. The positions of the trace for the third and the fourth readings are shown respectively by Fig. 12 (B) and Fig. 12 (C). Ordinates of subsequent dots are recorded in order by shifting the phase control in each instance until the dot falls on the vertical axis. Usually less than twenty readings are required to define the entire transition. Two ends are attained by following this procedure instead of by taking all readings with the trace in one position: (1) only a minimum amount of diligence is needed to insure recording *all* dots in the correct order and (2) errors are avoided which otherwise might be introduced as a consequence of a variation in vertical-deflection sensitivity by reason of horizontal deflection. Usually the responses for rising and for falling transitions are identical and only one need be recorded. An important exception occurs in the response of a detector for a vestigial-sideband system with high-percentage

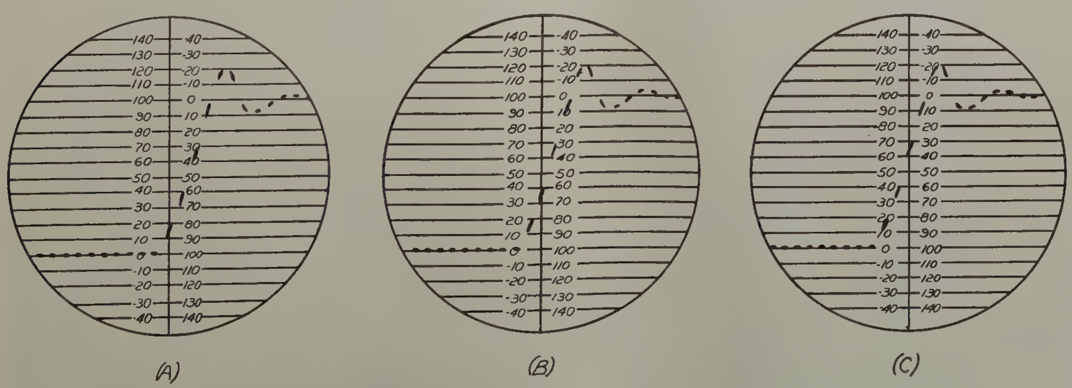


Fig. 12—(A) Position of square-wave trace for recording ordinate values of second dot. (Reading is 15.)
(B) Position for third dot. (Reading is 36.)
(C) Position for fourth dot. (Reading is 64.)

modulation.[5] The falling transition is recorded in the same manner as the rising transition except that the readings are made on the inverted scale beginning at the top. The measuring apparatus can be calibrated or checked by recording the output of the square-wave generator directly.

It is believed that portable square-wave measuring equipment of the type described will be found useful in experimental work, in routine checking, and in determining specifications of television apparatus. The recorded data are permanent and can be filed for future reference and comparison with similar square-wave records. Reconstruction of the square-wave response on paper is obtained readily by plotting the instantaneous recorded values. The form of the data is also directly suitable for analysis by the chart method described in a companion paper.[1]

[5] R. D. Kell and G. L. Fredendall, "Selective side-band transmission in television," *RCA Rev.*, vol. 28, pp. 425–444; April, 1940.

A New Direct Crystal-Controlled Oscillator for Ultra-Short-Wave Frequencies*

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Summary—*An ultra-high-frequency crystal oscillator is described which utilizes a mechanical harmonic of an AT or BT crystal. With the oscillator frequencies as high as 197 megacycles, harmonics as high as the 23rd have been excited. Taking the second electrical harmonic of the oscillator, frequencies as high as 300 megacycles, or 1 meter have been obtained. Since a mechanical harmonic is used, the crystal can be of a practical size to handle and adjust. The harmonic vibration of the AT and BT crystals have as low a temperature coefficient as the fundamental mode, and temperature coefficients of less than two parts per million per degree centigrade are easily obtained. Stability curves for this type of oscillator are shown and the results indicate that at 120 megacycles stabilities in the same order of magnitude as for ordinary crystal oscillators can be obtained. Without temperature or voltage control it appears likely that the frequency should remain constant to ±0.0025 per cent.*

Some measurements have been made of the properties of harmonic crystals at high frequencies. It was found that the Q of a crystal is independent of the frequency but in general increases with harmonic order. The ratio of capacitances r of a crystal increases as the square of the harmonic order. It is shown that in order to obtain a positive reactance in the crystal Q>2r. This relation will only be satisfied for harmonics of AT crystals less than the 7th. As a result oscillator circuits such as the Pierce circuit cannot be used to drive crystals at high harmonic frequencies. A discussion of oscillator circuits is given and it is shown that a capacitance-bridge oscillator circuit with the crystal in one arm is the best type to use for high-frequency harmonic crystals.

I. Introduction

DURING the last several years high-frequency vacuum tubes of moderate power output[1] have been developed which extend the commercially usable frequency spectrum into the ultra-short-wave region. Point-to-point station transmitters working at 120 megacycles and 150-megacycle aircraft transmitters are examples of such moderate power equipment. These applications require high-frequency oscillators with stabilities of the high order of magnitudes which are attainable with crystal oscillators. This stability has been obtained in the past by using crystal oscillators working under 20 megacycles with a number of stages of harmonic generation.

It is the purpose of this paper to describe a crystal oscillator which is controlled directly by a crystal resonance as high in frequency as 197 megacycles. This resonance is a mechanical harmonic of a low-coefficient AT-cut crystal. The mechanical harmonic has the same temperature coefficient as the fundamental AT crystal which can be made under two parts per million per degree centigrade. Since a mechanical harmonic is used the crystal is considerably thicker than it would be if it had to vibrate as its fundamental in the ultra-short-wave region, and hence it can be ground and adjusted much more easily than a very thin crystal. The particular oscillator whose properties are described here uses the 15th mechanical harmonic of an AT-cut crystal and produces a frequency of 120 megacycles.

Theoretically the electromechanical coupling of a harmonic crystal varies inversely as the order of the mechanical harmonic and the ratio of capacitances in the equivalent circuit varies proportionally to the square of the order of the mechanical harmonic. It becomes increasingly difficult to excite a harmonic vibration with the ordinary oscillator circuits and in fact harmonics higher than the 5th cannot usually be excited with the Pierce circuit for example. The reason for this is that the Pierce circuit requires the crystal to have a positive reactance in order that oscillation shall take place. It is shown below that in order for a positive reactance to occur in the crystal it is necessary that $k^2/2>1/Q$ or $Q>2r$ where Q is the ratio of reactance of the coil in the electrical representation of the crystal to its resistance, k the electromechanical coupling factor, and r the ratio of the shunt static capacitance C_0 of the

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[1] A. L. Samuel and N. E. Sowers, "A power amplifier for ultra-high frequencies," Proc. I.R.E., vol. 24, pp. 1464–1483; November, 1936; *Bell Sys. Tech. Jour.*, vol. 6, pp. 10–34; January, 1937.

crystal to the series capacitance C_1. The ratio of capacitances of a fundamental AT crystal as usually mounted is in the order of 1000 to 1 and hence $2r$ for the 5th harmonic would be 50,000. The Q of a crystal is usually not much larger than this so the 5th harmonic is about the highest harmonic that can be driven by the usual oscillator circuits.

For this application the circuit used consists of a high-frequency pentode with a tuned grid and plate

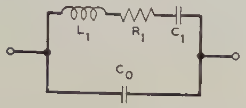

Fig. 1—Equivalent electrical circuit for piezoelectric crystal.

coupled back through a capacitance bridge of which the crystal forms one arm. For this circuit the crystal impedance does not have to have a positive reactance and it has been found possible to control the frequency of an oscillator with harmonics up to the 23rd or higher.

The stability obtainable with this oscillator is about the same as can be obtained with ordinary circuits at lower frequency. The stability with plate-voltage change is in the order of 0.05 cycle per megacycle per volt. The temperature coefficients of the crystal are under two parts per million per degree centigrade. Without regulation of voltage or temperature the frequency should be stable to ± 0.0025 per cent or better. With regulation the stability can be increased.

II. Properties of High-Frequency Crystals

In order that a crystal shall be useful in an ultra-high-frequency oscillator it is necessary that its Q shall remain high in the high-frequency range. Since no measurements have been published on how the Q's of crystals vary with frequency it was thought worth while to measure the Q's of AT-cut crystals over a

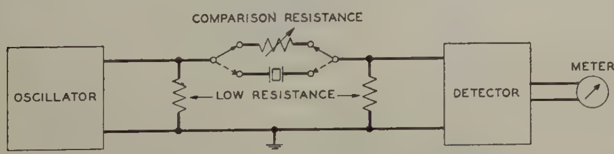

Fig. 2—Resistance method for measuring Q of crystal.

wide frequency range. For this purpose 9 crystals were obtained whose fundamentals varied in frequency from 800 kilocycles to 20 megacycles. The Q of a crystal is defined with respect to the equivalent circuit of the crystal shown in Fig. 1. Here the static capacitance of the crystal is designated by C_0, while the effect of the motional impedance of the crystal is represented by the series-resonant circuit L_1, C_1, and R_1. The Q of the crystal is defined as the ratio $(2\pi f_R L_1)/R_1$ where f_R is the resonant frequency.

Three independent methods were used to measure the Q of the crystals. At the lower frequencies the method used was the one described in a former paper[2] and shown in Fig. 2. It consists in measuring the resonant frequency f_R, the antiresonant frequency f_A, the resistance at resonance R, and the static capacitance C_0 of the crystal. From these the Q of the crystal can be calculated from the formula

$$Q = \frac{f_R^2/(f_A^2 - f_R^2)}{2\pi f_R C_0 R} \doteq \frac{1}{4\pi R C_0 \Delta} \quad \text{where } \Delta = f_A - f_R. \tag{1}$$

At higher frequencies, the measurement of resistance becomes less reliable. At these frequencies, however, the reading of vacuum-tube voltmeters is satisfactory, so the circuit of Fig. 2 was modified to that of Fig. 3,

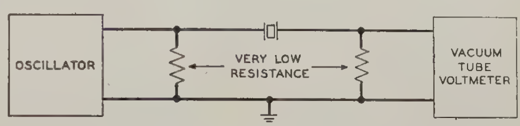

Fig. 3—Voltmeter method for measuring Q of crystal.

and the resonant and antiresonant frequencies were measured as well as the voltage across the low terminal resistance at resonance and antiresonance. It is readily shown that the Q of the crystal will be given by the formula

$$Q = \frac{f_R}{2\Delta}\sqrt{\frac{V_R}{V_A}} \quad \text{where} \quad \Delta = f_A - f_R \tag{2}$$

and V_R and V_A are the voltages across the terminating resistance at the resonant and antiresonant frequencies, respectively. A third method used was to measure the voltage across the terminating resistance at antiresonance and a few cycles from antiresonance. For this case the Q of the crystal is given by

$$Q = \frac{f_A}{\Delta}\left[\left(\frac{V_C}{V_A}\right)^2 - 1\right] \tag{3}$$

where Δ is the difference in frequency between the antiresonance and the frequency for which the voltage V_C is measured. All three methods were compared on several crystals and checked within a few per cent.

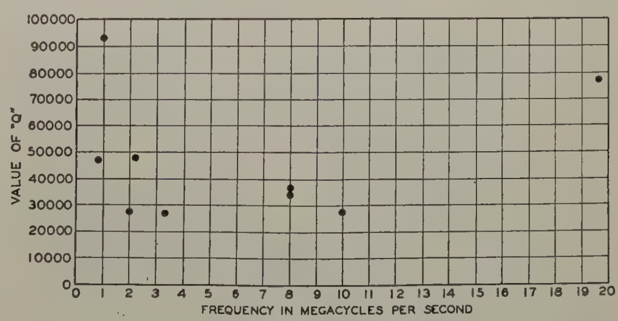

Fig. 4—Measurements of the Q of fundamental AT-cut crystals over a frequency range.

The measurement of the fundamental frequencies of the 9 crystals are shown in Fig. 4. Although the Q varies over a considerable range from crystal to crystal,

[2] W. P. Mason, "Electric wave filters employing crystals as elements," *Bell Sys. Tech. Jour.*, vol. 13; pp. 405–452; July, 1934.

there is no significant trend with frequency. The ratio of capacitances for the fundamental AT crystals was around 1000 to 1.

The harmonic vibrations of three of the crystals were also measured and the resulting Q's are shown in Fig. 5. The crystal having a fundamental at 1 megacycle has a fundamental Q of 98,000 and progressively

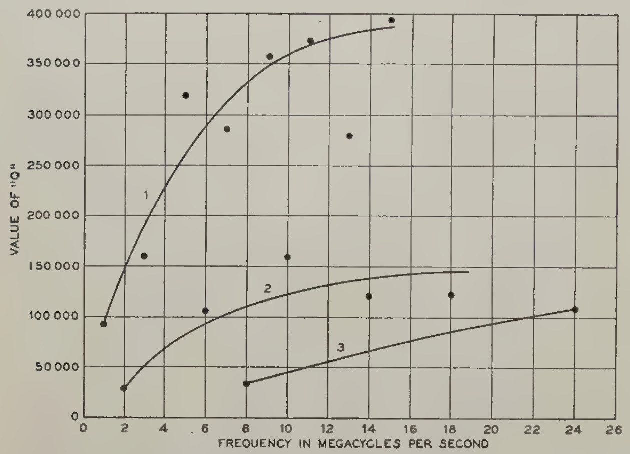

Fig. 5—Measurements of the Q of harmonic AT-cut crystals.

higher Q's at the higher-order harmonics. The other crystals have lower initial Q's but all show an increasing Q with harmonic order. The crystal having a fundamental of 8 megacycles is one of the crystals used in the oscillator described in this paper.

It is thought that the variation in Q of shear-vibrating crystals is due to very minute cracks on the surface which cause a small amount of rubbing when the crystal vibrates. This is borne out by the fact that etched or polished crystals usually have a higher Q than nontreated crystals. This is probably also the explanation of the higher Q in the harmonic crystals for a harmonic crystal is similar to a number of crystals in series with a number of vibrating surfaces as shown in Fig. 6. All the internal surfaces will be whole and free from cracks. If most of the resistance is associated with the outside faces it follows that a harmonic crystal with a large number of internal faces should have less dissipation than a fundamental crystal. At the high-order harmonic the Q seems to approach an asymptote of 400,000 to 500,000, which is probably the Q associated with the internal dissipation.

These data indicate that as far as internal dissipation is concerned, the Q of a crystal is independent of the frequency. If, however, one goes from a condition for which the Q is controlled by the surface dissipation to a condition where the Q is controlled by the internal dissipation, the Q will increase.

A measurement of the ratio of capacitances of the higher-order harmonics indicated that the ratio of capacitances increases about as the square of the harmonic order. This is in agreement with theory for in an odd-harmonic crystal of order n, only $1/n$th of the piezoelectric effect is engaged in driving the crystal at its harmonic frequency and the other $(n-1)/n$ parts annul each other. The electromechanical coupling factor is[3]

$$k = d_{26_n'}\sqrt{\frac{4\pi c_{66'}}{K}} \tag{4}$$

where $d_{26_n'}$ the piezoelectric constant is d_{26}/n, c_{66} is the shear elastic constant, and K the dielectric constant of the crystal. It follows that the coupling varies inversely as the harmonic order. The ratio of capacitances r being[3]

$$r = \frac{8}{\pi}\left(\frac{1-k^2}{k^2}\right) \tag{5}$$

is approximately proportional to n^2.

It is a matter of interest to find how the impedance of a harmonic crystal varies with the harmonic. It has been shown previously[3] that the impedance of a piezoelectric crystal is given by the expression

$$Z_c = \frac{-j(1-k^2)}{\omega C_0}\left[\frac{1 - f^2/f_1^2 + \dfrac{j}{Q(1-k^2)}}{1 - f^2/f_2^2 + j/Q}\right] \tag{6}$$

where C_0 is the static capacitance of the crystal, f_1 the resonant frequency, and f_2 the antiresonant frequency. Also, $f_1^2 = f_2^2(1-k^2)$. Inserting this value and expressing the equation in the form of a resistance and reactance we find

$$Z_c = \frac{\dfrac{k^2}{Q} - j\left[1 - k^2 - \dfrac{f^2}{f_2^2}(2-k^2) + \dfrac{f^4}{f_2^4} + \dfrac{1}{Q^2}\right]}{\omega C_0\left[\left(1 - \dfrac{f^2}{f_2^2}\right)^2 + \dfrac{1}{Q^2}\right]}. \tag{7}$$

Fig. 6—Harmonic mode of motion.

The reactance term will have a maximum when $f^2 = f_2^2(1-1/Q)$. At that frequency the impedance of the crystal will be

$$\frac{\dfrac{k^2Q}{2} + j\left(\dfrac{k^2Q}{2} - 1\right)}{\omega C_0}. \tag{8}$$

[3] W. P. Mason, "An electromechanical representation of a piezoelectric crystal used as a transducer," Proc. I.R.E., vol. 23, pp. 1252–1264; October, 1935.

Hence, in order that the crystal shall exhibit a positive reactance

$$\frac{k^2Q}{2} > 1 \quad \text{or} \quad k^2 > \frac{2}{Q}. \tag{9}$$

But the ratio of capacitance is given by (5). Hence, the condition[4] for the crystal to exhibit a positive reactance is

$$Q > 2(1 + r) \doteq 2r. \tag{10}$$

If we assume that Q is in the order of 75,000 or less, which is certainly higher than can be obtained in most untreated crystals, and a ratio r of 1000 for a fundamental AT crystal, the highest-order harmonic that will show a positive reactance is the 5th harmonic for an AT-cut crystal.

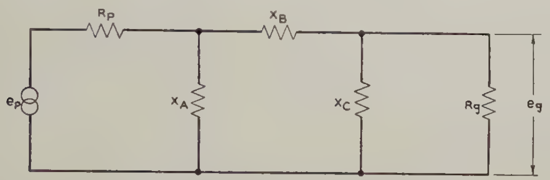

Fig. 7—Equivalent circuit of Pierce-type oscillator.

III. Standard Oscillator Circuits for Crystals

As a consequence of the fact that high-harmonic crystals will not show a positive reactance, it follows that the standard oscillator circuits, such as the Pierce circuit, cannot be used to excite high-harmonic-order crystals. The only type of oscillator circuit which will annul the large shunt capacitance of the crystal, which prevents the crystal from having a positive reactance near the resonant frequency, is a bridge-type circuit, as discussed in Section IV. When this reactance is annulled, however, the resulting circuit reduces to a circuit similar to the Pierce circuit or to one in which the phase is reversed by the use of a phase-reversing transformer, depending on whether the crystal is in the series or lattice arm of the bridge. Since the stability of the circuit depends on which connection is used and can be evaluated from a consideration of the stability of standard oscillator circuits, it has appeared worth while to include a discussion of them.

The Pierce-type oscillator circuit consists of a π-network connection of reactances between the plate and grid of the tube as shown in Fig. 7. Such a circuit is required to produce a 180-degree phase shift in order to offset the 180-degree phase shift produced in the vacuum tube. The other common connection is to use a phase-reversing transformer or two tubes in tandem, so that the network itself is required to produce a zero phase shift.

If we consider a tube as represented by the circuit shown in Fig. 7, following Llewellyn,[5] the condition for zero gain and zero phase shift are given by

$$\frac{R_p}{R_g} = -\left[\frac{X_A}{X_A + X_B}\right]\left[\mu + \frac{X_B + X_C}{X_C}\right] \tag{11}$$

and

$$R_pR_g = \frac{X_AX_BX_C}{X_A + X_B + X_C} \tag{12}$$

where μ is the amplification factor of the tube, R_p and R_g are the plate and grid resistances of the tube, and X_A, X_B, X_C the coupling reactances which incorporate the distributed capacitances of the tubes. These formulas assume that the tube introduces a phase angle of 180 degrees. In case two tubes are used in tandem or the 180-degree phase shift is produced in a unity-coupled transformer, the sign of μ in (11) will be reversed. For two tubes, μ will have the significance of the ratio between the voltage applied in the plate circuit of the last tube to the voltage applied on the grid of the first tube.

For the first type circuit, μ will ordinarily be greater than $(X_B+X_C)/X_C$ and hence X_B will have to be of opposite sign to X_A and greater in absolute magnitude; for writing,

$$\frac{X_A}{X_A + X_B} = -m \text{ we have } X_B = -\frac{(1+m)}{m}X_A. \tag{13}$$

Introducing this value into (12), we have

$$R_pR_g = -\frac{X_A^2(1+m)}{m - \dfrac{X_A}{X_C}}. \tag{14}$$

Since X_A^2 is positive, the right-hand side of (14) can only be made positive if X_A is the same sign as X_C and

$$mX_C > X_A. \tag{15}$$

These equations show that X_A and X_C must be of the same sign and opposite to X_B while the sum of X_A and X_C must be larger in magnitude than X_B. If $(X_B+X_C)/X_C$ is of opposite sign to μ and greater in magnitude it is readily shown that the resulting condition on X_A will not satisfy (13) and hence the oscillator will not oscillate under these conditions.

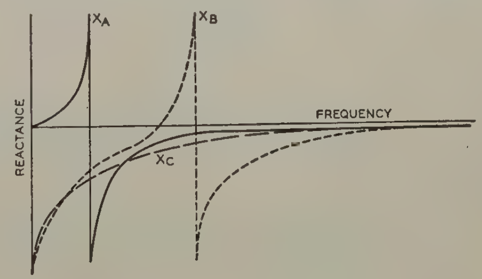

Fig. 8—Reactance curves for Pierce-type oscillator.

Since we do not know R_p and R_g explicitly, the frequency cannot be calculated definitely from these

[4] This condition was first derived by R. A. Sykes.
[5] F. B. Llewellyn, "Constant frequency oscillators," Proc. I.R.E., vol. 19, pp. 2063–2094; December, 1931.

equations unless R_p, R_g, and the distributed capacitances of the tubes are evaluated as functions of the voltage and current conditions in the tube. Ways for minimizing the variation with tube and circuit conditions, however, are evident from (12). If, for example, the crystal is connected in the X_B arm, it is

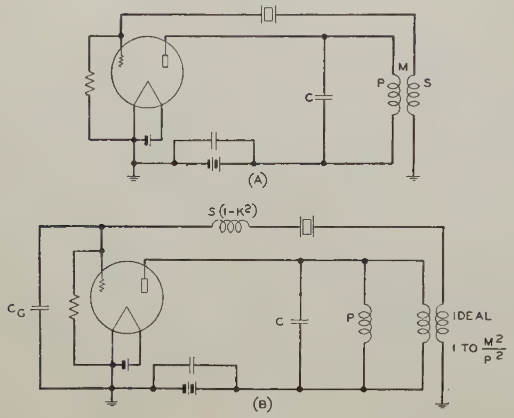

Fig. 9—Oscillator using phase-reversing transformer.

readily shown that the maximum voltage and tuning stability will occur when the X_A arm, which usually consists of a tuned circuit, is tuned for a frequency considerably lower than the crystal resonance. Under these conditions, as shown by Fig. 8, X_A is much smaller than X_B or X_C, the value of R_g is large, and the frequency is determined nearly by the condition that $X_A+X_B+X_C=0$. A change in R_pR_g caused by a change in voltage will produce only a small frequency change on account of the large value already existing for the product R_pR_g. The maximum stability for this type of oscillator circuit occurs for a minimum output.

On the other hand, if we employ a circuit with two tubes or a phase-reversing transformer, the point of greatest voltage stability can be made to come at the highest power output. One such circuit is shown in Fig. 9(A). For a finite coupling between the two transformer windings, the equivalent network of the circuit is shown in Fig. 9(B). The leakage reactance can be combined with the crystal reactance if desired, in which case it will lower the resonant point slightly or if desired a series capacitance can be inserted to annul the leakage reactance.

The condition for oscillation for this type of circuit can be obtained from (11) and (12) by reversing the sign of μ in (11). If we let

$$\frac{X_A}{X_A+X_B}=m_1 \quad \text{or} \quad X_B=X_A(1-m_1)/m_1 \qquad (16)$$

then, in order to oscillate, m_1 must be positive. If m_1 is between zero and unity then

$$X_A=(m_2-m_1)X_C \text{ where } 0<m_2<\frac{m_1(\mu-m_1)}{1-m_1}. \qquad (17)$$

If m_1 is greater than unity,

$$\frac{-m_1(\mu+m_1)}{m_1-1}<m_2<0.$$

This type of circuit will have great stability when the crystal is worked near its resonance frequency for which case X_B is a small quantity Δ. Then, for this case,

$$\frac{R_p}{R_g}=\mu-1;\quad R_pR_g=\Delta\left(\frac{X_AX_C}{X_A+X_C}\right). \qquad (18)$$

If we let X_A and X_C antiresonate at the crystal resonant frequency f_R, the product of Δ by the antiresonant impedance will be finite and can be made equal to R_pR_g. Under these conditions a large change in R_pR_g will not change the frequency since no phase shift requiring a shift in frequency will occur. This analysis does not take account of the change in dielectric constant with voltage for the tube capacitances entering the circuit but it has been found experimentally that this effect is small and does not alter appreciably the maximum stability conditions. Even when dissipation is associated with all of the elements, this relation is still true as can be demonstrated with reference to Figs. 10(A) and 10(B).

As shown by (18), the ratio of $R_p/R_g=\mu-1$, which is the maximum value possible. Hence, the output will be greatest at the voltage-stabilized point also. Since R_g will be considerably smaller than R_p an increase in output can be obtained by making the ratio of the transformer output impedance to its input impedance

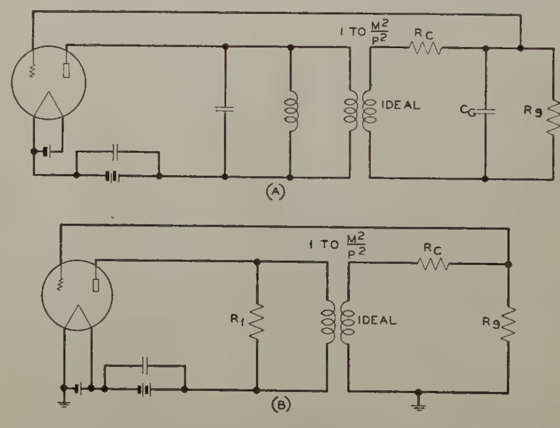

Fig. 10—Impedances at crystal resonance for phase-reversing oscillator.

less than unity. For this case the frequency and amplitude equations become

$$\frac{R_p}{R_g}=\frac{X_A}{\phi^2X_A+X_B}\left[\mu\phi-\frac{X_B+X_C}{X_C}\right]$$

$$R_pR_g=\frac{X_AX_BX_C}{\phi^2X_A+X_B+X_C} \quad \text{or} \qquad (19)$$

$$R_pR_g\phi^2=\frac{X_A\phi^2X_BX_C}{\phi^2X_A+X_B+X_C}$$

where ϕ^2=impedance transformation ratio$=M^2/P^2$, where M is the mutual inductance of the transformer, and P the primary inductance. For X_B very small or for oscillations near the resonant frequency

$$\frac{R_p}{R_g}=\frac{\mu\phi-1}{\phi^2}; \quad R_pR_g\phi^2=\left(\frac{X_A\phi^2X_C}{\phi^2X_A+X_C}\right)\Delta. \tag{20}$$

Hence, the ratio of R_p/R_g will be a maximum and consequently the output will be a maximum when $\phi=2/\mu$. If the resistances of the elements, as represented by

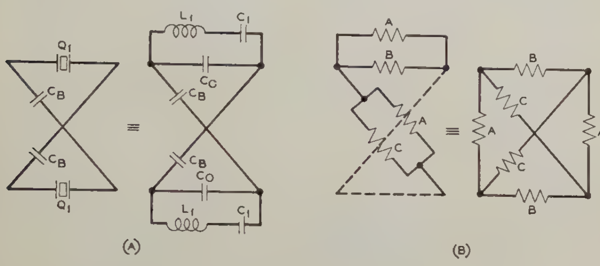

Fig. 11—Use of crystal in bridge circuit.

Fig. 10(B), are taken account of this transformation ratio for maximum output will be less. The stability with reactance tuning will also be increased by using a step-down transformer since the impedance of the crystal will effectively be increased compared to the impedance of the tuning reactance. Hence, it requires a larger change in tuning of the plate reactance to cause a given frequency change in the oscillator.

IV. Oscillator Circuits for Use with High-Frequency Harmonic Crystals

Neither of these two standard types of circuits can be used directly with high-harmonic-type crystals on account of the fact that the sign of the reactance does not become positive in the resonance region. They could be used to control the frequency by using the crystal as a negative reactance, but in that case it is difficult to locate the control in the resonance region. We note, however, on examining the equivalent circuit of a crystal shown in Fig. 1, that the reason the reactance does not go positive is that the series-resonant arm is shunted by a very large condenser. If this condenser could be neutralized, the remaining impedance arm would consist of a simple tuned circuit which changes from negative to positive reactance at the resonant frequency.

In crystal-filter work[6] it has been shown that the static capacitance of a crystal can be balanced out by incorporating the crystal in a lattice network of capacitances as shown in Fig. 11(A). The drawing shows two crystals having the same constants but they can be represented actually by a single crystal with two sets of plates. By virtue of the network theorem shown in Fig. 11(B), the static capacitances of the crystal together with the balancing condenser C_B can be removed to the end of the lattice leaving a network of reactances. In this network the series arms are the resonance series arms of the equivalent representation of the crystal, and the shunt arms are the static capacitance of the crystal.

Suppose now that we incorporate this crystal lattice in the oscillator network shown in Fig. 12(A). Then as before we can take out the shunt capacitance of the crystal and the balancing capacitance to the ends of the network leaving the circuit shown in Fig. 12(B). Assuming perfect coupling in the transformer, the shunt capacitance C_B can be joined to C_0 to tune the input coil, and C_B can be joined to C_2 to tune the output coil. If we have 180-degree phase reversal in the transformer, and tune the two end coils to antiresonance at the resonant frequency of the crystal, the phase shift around the feedback path will be 360 degrees and the oscillator will oscillate at the crystal resonant frequency, which, as discussed in Section III, will result in the greatest stability at the frequency of maximum output In case the transformer gives no change in phase, the crystal has to be put in the lattice arm to give this condition. In case the coil coupling is not unity, a small leakage-reactance coil appears. This can be annulled at the resonant frequency of the crystal by a small series capacitance. Hence, the resultant reactance at the resonant frequency will be the same as that shown in Fig. 12(B).

The crystal network as shown in Fig. 11 has two identical crystals or a crystal with two sets of plates. This is rather objectionable at high radio frequencies, since it is difficult to divide the plates for very small

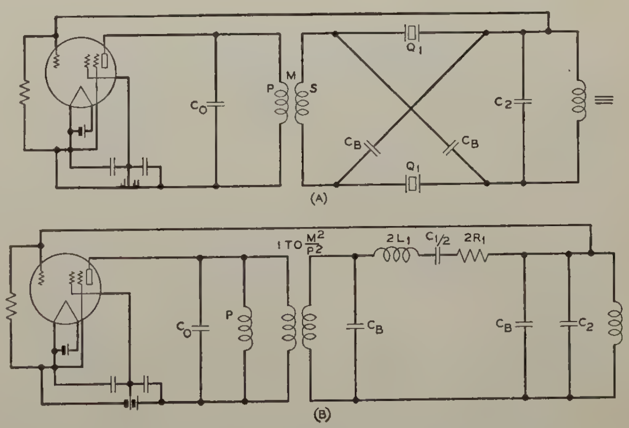

Fig. 12—Unbalanced oscillator incorporating bridge circuit with balanced crystal.

crystals. Fortunately this is not necessary for we can show that the crystal lattice with a single crystal shown in Fig. 13(A) is equivalent to the balanced lattice with two crystals. This follows from the network equivalence of Fig. 13(B), which states that the series arm of the balanced lattice is equal to

$$\frac{B(A+C)+2AC}{2B+A+C}=D. \tag{21}$$

[6] See W. P. Mason, "Resistance compensated band-pass crystal filters for unbalanced circuits," *Bell Sys. Tech. Jour.*, vol. 16, pp. 423–436; October, 1937.

The crystal impedance C is given by (6) while A and B are

$$A=\frac{-j}{\omega C_A}; \qquad B=\frac{-j}{\omega C_B}. \tag{22}$$

Inserting these values in (21), we find the equivalent impedance D to be given by

$$D=\frac{-j(1-k'^2)}{C_0'\omega}\left[\frac{1-\dfrac{f^2}{f_2'^2(1-k'^2)}+\dfrac{j}{Q'(1-k'^2)}}{1-\dfrac{f^2}{f_2'^2}+\dfrac{j}{Q'}}\right] \tag{23}$$

where

$$C_0'=\frac{C_0(2C_A+C_B)+C_AC_B}{C_0+C_A+2C_B} \tag{24}$$

$$f_2'^2=f_2^2\left[1-\frac{k^2(C_AC_B)}{C_0(2C_A+C_B)+C_AC_B}\right];$$

$$Q'=Q\left[1-\frac{k^2(C_AC_B)}{C_0(2C_A+C_B)+C_AC_B}\right]$$

$$k'^2=\frac{2k^2C_0(C_A+C_B)^2}{(C_0+C_A+2C_B)\left[C_0(2C_A+C_B)+C_AC_B-k^2C_AC_B\right]}.$$

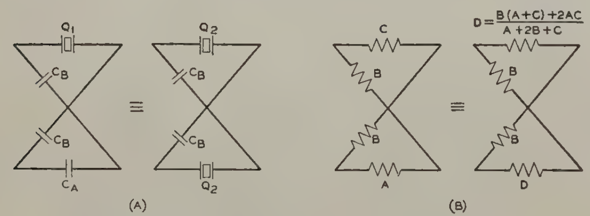

Fig. 13—Equivalence between balanced and unbalanced lattice networks.

Hence, the impedance of the D arms is that of a crystal with somewhat different constants than that for the original single crystal. If

$$C_0C_A=C_B^2 \tag{25}$$

the values in the above equations become

$$C_0'=C_B;$$

$$f_2'^2=f_2^2\left[1-\frac{k^2C_A^2}{(C_A+C_B)^2}\right];\quad Q'=Q\left[1-\frac{k^2C_A^2}{(C_A+C_B)^2}\right]$$

$$k'^2=\frac{2k^2C_AC_B}{(C_A+C_B)^2-k^2C_A^2}\doteq\frac{2k^2C_AC_B}{(C_A+C_B)^2}. \tag{26}$$

If C_A is nearly equal to C_B, which will be the ordinary condition of operation, $f_2'^2=f_2^2(1-k^2/4)$; $Q'=Q(1-k^2/4)$; $k'^2=k^2/2$.

Since the shunting capacitance of the crystal C_0' is equal to C_B, all of the capacitances C_B can be removed from the lattice arms to the ends of the circuit leaving the equivalence shown in Fig. 14, in which the series-resonant arm representing the crystal impedance has four times the impedance of the single crystal used in the lattice. The oscillator shown in Fig. 14 is then capable of driving a crystal at a high harmonic and of being controlled by the resonance characteristic of the crystal at that harmonic. A 180-degree phase shift is introduced by the transformer and the resonant circuits are tuned so that their antiresonant frequencies

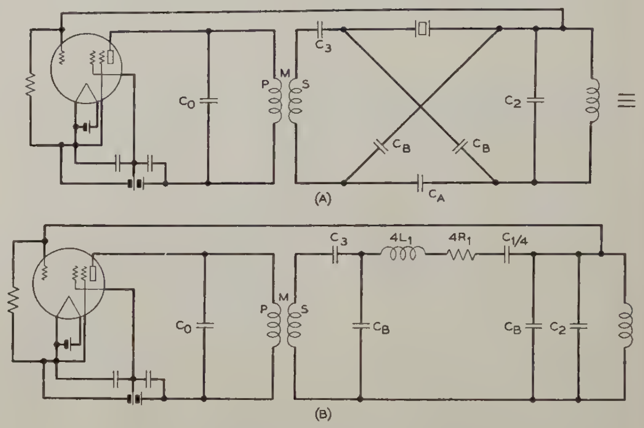

Fig. 14—Unbalanced oscillator incorporating unbalanced crystal bridge.

coincide with the resonant frequency of the crystal. An alternative arrangement is to use a transformer with no change of phase and the crystal in a lattice arm. This is the condition for maximum output and maximum stabilization against voltage changes as pointed out previously.

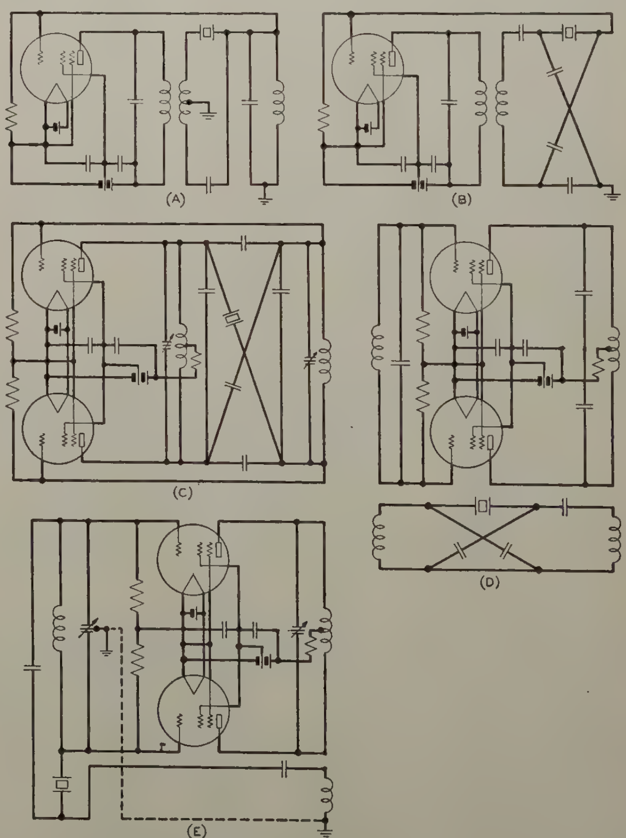

Fig. 15—Balanced and unbalanced oscillators.

Other circuits for which the same process can be applied are shown in Fig. 15. Some of these circuits are unbalanced and some balanced. At ultra-high frequencies it is often desirable to use balanced circuits

and double pentodes[1] such as the 240H have been developed for this purpose. For such systems it is desirable to use balanced oscillator circuits, and most of the experimental work recorded here has been done with the circuits of Fig. 15(C) and Fig. 15(E). Fig. 15(C) is especially advantageous for the crystal and all

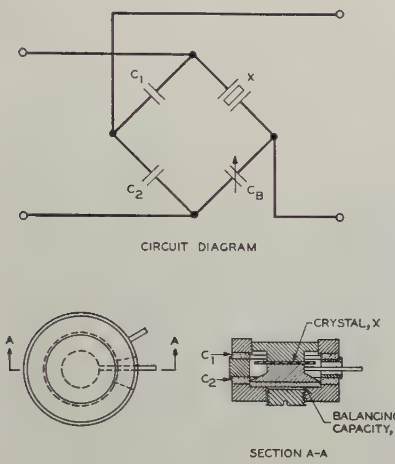

Fig. 16—Crystal bridge.

of the balancing capacitors can be put in one self-contained structure, as shown in Fig. 16, which can be enclosed in a metal vacuum-tube holder and evacuated to eliminate the effects of air damping and humidity on the crystal and bridge balance. The dielectric employed is usually fused quartz on account of its low loss and similar temperature variation of dielectric constant to that of crystalline quartz. In some cases series condensers have been placed on each side of the bridge to control the amount of feedback. The variable

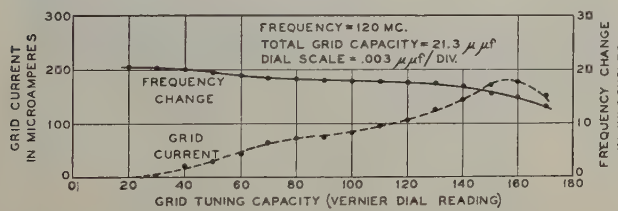

Fig. 17—Effect of grid tuning on frequency of oscillator.

tuning condensers are sometimes made with one component having a negative temperature coefficient to compensate for the positive coefficient of the metal tuning coils. This insures that the oscillator gain will remain at its maximum value over a wide temperature range.

V. Experimental Results

In the experimental results obtained, the circuit of Fig. 15(C), with a self-contained bridge, has usually been employed since it was desirable to work the oscillator into a balanced 240H pentode amplifier. In order to make the crystal control at high radio frequencies it is necessary to employ a circuit with as much gain as possible at the high frequencies. This indicates that high-frequency pentodes should be employed; accordingly, 954 acorn pentodes were used.

With this arrangement, it was found possible to drive crystals at harmonics as high as the 23rd and frequencies as high as 197 megacycles or 1.5 meters wavelength. Furthermore, by taking the second electrical harmonic of the oscillator, a frequency of 300 megacycles or 1 meter was obtained with good output. This probably does not represent the upper limit for by using circuits with more gain the loss inserted by the crystal circuit can be overcome and the oscillator be crystal controlled to even higher frequencies.

The adjustment of this oscillator is not difficult. The crystal bridge is left unbalanced first and the oscillator will oscillate uncontrolled by the crystal. Its frequency is determined by the resonance of the coil-condenser systems on the ends. The frequency is adjusted to a somewhat lower frequency than that of the desired mode of the crystal and the coils are tuned for maxi-

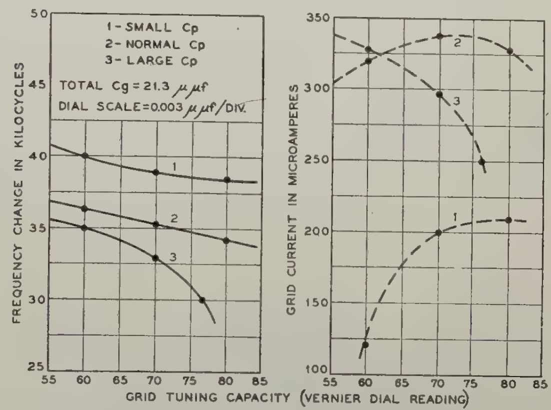

Fig. 18—Effect of grid and plate tuning on frequency of oscillator.

mum output at this frequency. The crystal bridge condenser is then adjusted towards balance and the oscillations will usually stop. The grid and plate coils are then tuned to the crystal frequency and the oscillator will then be controlled only by the crystal.

Some stability curves for such an oscillator were obtained for a frequency of 120 megacycles or 2.4 meters. The output obtained was 80 volts across the output impedance of 25,000 ohms when the plate voltage was 250 volts and the screen voltage 100 volts. The frequency variation with grid tuning for one crystal used is shown in Fig. 17. The grid current obtained is also plotted on this curve. As can be seen the maximum variation is in the order of 5 kilocycles from the point of maximum grid current to the point where oscillations ceased. The curves for plate tuning are similar to those for grid tuning. The over-all frequency change for both plate and grid tuning is shown in Fig. 18. This over-all change amounts to 10 to 12 kilocycles or about 100 parts in a million. This range represents the controllable range of the crystal and, therefore, the maximum frequency deviation which would occur for any possible combination of circuit changes unless the bridge becomes unbalanced and the circuit oscillates uncontrolled. Since in the extreme tuning positions the output is very low, the possibility of operating this far

off frequency is not great, and the usable range is probably not over one half this or ± 0.0025 per cent.

The change in frequency with plate voltage for this type of circuit is quite small for the maximum output. The reason for this is pointed out in the section on standard oscillators. When the circuit is tuned some-

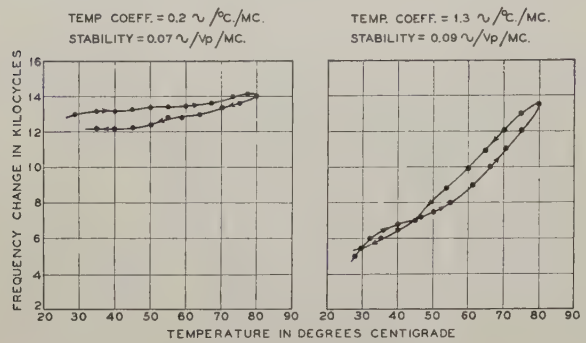

Fig. 19—Frequency change as a function of temperature.

what differently, the variations may amount to 0.05 to 0.1 cycle per megacycle per volt.

The temperature-frequency curves for two AT plates vibrating at their 15th harmonic are shown in Fig. 19. The average coefficient over a 50-degree-centigrade temperature range was 1.3 parts in 10^6 per degree centi-

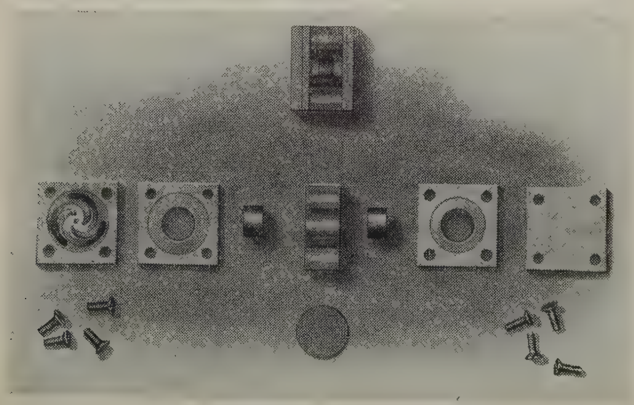

Fig. 20—Photograph of crystal and holder.

grade for one and 0.2 part for another. Five crystals were tested and all had a coefficient less than 2 parts in 10^6 per degree centigrade. One remarkable feature of these crystals is brought out by the temperature curves; that is, no hops in frequency were experienced in any of the temperature runs. This appears to be due to the fact that the coupling with harmonics of the low-frequency modes decreases with decreasing ratio of thickness to the diameter of the plate. This crystal was 0.21 millimeter in thickness and 12 millimeters in diameter and operated on the 15th harmonic, hence the ratio of diameter to effective thickness is 12 to 0.014 or 860 to 1, and the unwanted couplings are practically zero. Near-by frequencies may still be found if the plate is not exactly flat but they are of the same type and have the same temperature coefficient, with the result that they do not come nearer the desired resonance and hence cause little trouble.

It is important for this work to have the crystal of uniform thickness in order to obtain a large output and a single response at the desired harmonic frequency. They are polished and the degree of flatness determined by interference fringes. A photograph of the crystal and holder is shown in Fig. 20. In this figure some interference fringes are observable which snow the degree of flatness obtained. It is difficult to extend

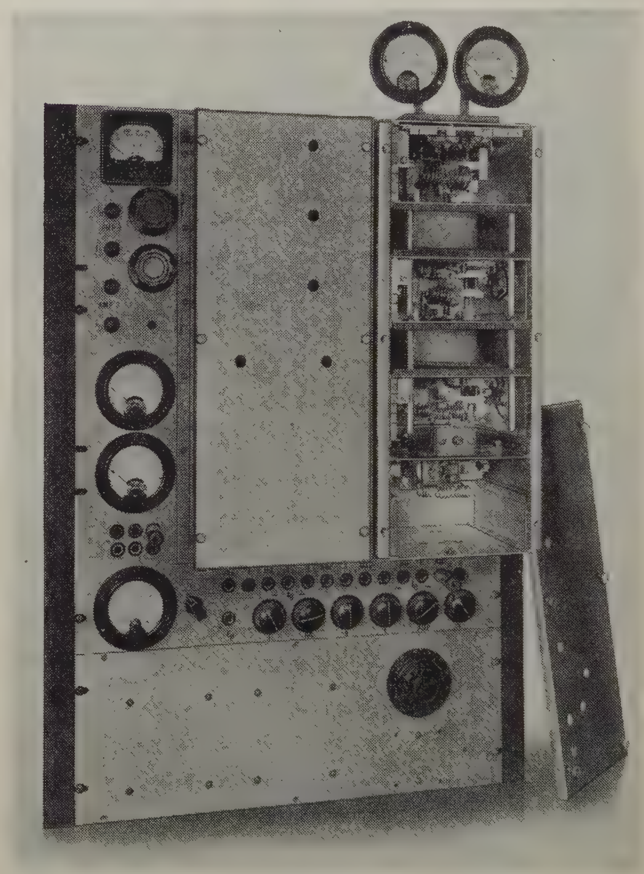

Fig. 21—Photograph of crystal oscillator and amplifier.

the flat region out to the edges since there is a tendency for the crystal to become convex. For this reason the diameter of the crystals is made twice that of the electrode area. The holder itself has very flat electrodes and is designed to center the electrodes on the crystal thus utilizing the most nearly flat portions of the crystal.

A crystal oscillator of this sort was used in conjunction with two 240H amplifying pentodes to deliver 12 watts output to the antenna at 120 megacycles. A photograph of the equipment is shown in Fig. 21. After $5\frac{1}{2}$ hours of continuous running the frequency had changed about 800 cycles or 7 parts in 10^6. Day-to-day checks indicated frequency agreement within several hundred cycles. Continuous runs indicate that the frequency variation under ordinary operating conditions should not be more than ± 25 parts in a million or ± 3 kilocycles at 120 megacycles.

An Evaluation of Radio-Noise-Meter Performance in Terms of Listening Experience*

CHARLES M. BURRILL†, MEMBER, I.R.E.

Summary—*An account is given of listening tests conducted, with the co-operation of the Joint Co-ordination Committee on Radio Reception of the Edison Electric Institute, National Electrical Manufacturers Association, and Radio Manufacturers Association, for the purpose of indicating how closely instruments made in accordance with the latest radio-noise-meter specifications of the Joint Co-ordination Committee meet the objective of giving readings proportional to annoyance for all types of radio noise. Thirty people participated in the tests which involved three types of radio noise and three different radio-noise meters. Standard statistical methods are used in analyzing the results, and these methods are explained in simple fashion for the benefit of radio engineers who are unfamiliar with statistical science. The general conclusion is that the new radio-noise-meter performance is very satisfactory.*

Introduction

A MAJOR objective of radio-noise-meter design is to provide an instrument which will give, for all kinds of radio noise, indications which are proportional to the annoyance factor or nuisance capability of the noise. This annoyance factor must ultimately be determined by listening experience. Therefore, the performance of each new design should be calibrated or evaluated by means of listening tests, in which the instrument readings are compared with the judgments of a jury of listeners as to the vexatiousness of the noises measured. The general problems involved in this basic procedure of radio-noise measurement have been discussed at length elsewhere.[1,2] The principal purpose of the present paper is to describe a series of tests which were made to evaluate, in terms of listening experience, the performance of radio-noise meters built in accordance with the latest specifications of the Joint Co-ordination Committee on Radio Reception of the Edison Electric Institute, National Electrical Manufacturers Association, and Radio Manufacturers Association.[3] A further purpose is to analyze and state the results of these tests in simple, easily understood terms in conformance with standard statistical practice. The exposition will therefore serve to illustrate the application of statistical methods to a radio-engineering problem, and to suggest to radio engineers relatively unfamiliar with such methods how they might be more widely used.

The tests to be described were planned and carried out, with the co-operation of the Joint Co-ordination Committee, by Dudley E. Foster at the RCA License Laboratory in New York City. Two manufacturers of radio-noise meters loaned the instruments which were used. The tests were held on the morning of May 24, 1940, preceding a regular meeting of the Joint Co-ordination Committee, so that the committee members and their guests could form the jury of listeners.

The Radio-Noise Meters

Three instruments were used, designated A, B, and C. A and B were of different manufacture but both were intended to be in accordance with the Joint Co-ordination Committee specifications.[3] C was similar to A, but incorporated time constants of 1 and 160 milliseconds, such as were being considered for standardization in Canada, instead of the time constants of 10 and 600 milliseconds adopted by the Joint Co-ordination Committee.

The Radio-Noise Sources

The noise from a direct-current commutator-type motor, that from an electric razor of the vibrator type, and the clicks from a direct-current relay were chosen for use as typical of a large percentage of radio noises encountered in broadcast reception. These noise sources covered the range from fairly continuous noise to impulse noise of repetition rates as slow as about one per second.

The Program Source

The program of one of the National Broadcasting Company networks, obtained by direct wire from the studio, was used to modulate a standard-signal generator, which in turn was used to excite a conventional

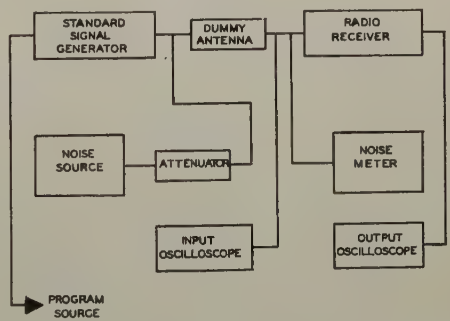

Fig. 1—Block diagram of the apparatus used in the tests.

broadcast receiver through a dummy antenna. The noise was also applied, through an attenuator, to the dummy antenna. A block diagram of the arrangement is shown in Fig. 1. The oscilloscopes were used in making preliminary adjustments and to indicate to the listeners, prior to the tests, what the noise waveforms were like. They were not used during the actual

* Decimal classification: R270. Original manuscript received by the Institute, December 10, 1941. Presented, Annual Convention, New York, N. Y., January 9, 1941.

† RCA Manufacturing Co., Inc., Camden, N. J.

[1] C. J. Franks, "The measurement of radio noise interference," *RMA Eng.*, vol. 3, pp. 7–10; November, 1938.

[2] C. M. Burrill, "Progress in the development of instruments for measuring radio noise," Proc. I.R.E., vol. 29, pp. 433–442; August, 1941.

[3] "Methods of measuring radio noise, a report of the Joint Co-ordination Committee on Radio Reception of the Edison Electric Institute, National Electrical Manufacturers Association, and Radio Manufacturers Association." Edison Electric Institute Publication No. G9, National Electrical Manufacturers Association Publication No. 107, Radio Manufacturers Association Engineering Bulletin No. 32; February, 1940.

tests. Two different receivers were used, a 13-tube console model and a 6-tube table model. The program was taken just as it came: speech, music, announcements, and all.

Quality of Reception

The quality of the program reception, with respect to noise interference, was rated by each listener in accordance with the following scale or code devised by the author some time ago:[4]

A—Entirely satisfactory
B—Very good, background unobtrusive
C—Fairly satisfactory, background plainly evident
D—Background very evident, but speech easily understood
E—Speech understandable only with severe concentration
F—Speech unintelligible

Intermediate grades were recognized, for example, B− was considered identical with C+ and half way between B and C.

This scale is a composite one, intended to be of use for programs consisting of either speech or music, whether listened to for entertainment as, for example, symphonic music, or for intelligence as, for example, news broadcasts. It is recognized that there is some loss of precision in using a single scale for all types of program. However, in the present state of the art of radio-noise measurement, the refinement of separate evaluations for different program types does not appear to be warranted. In using the present scale it is found that the three highest grades, A, B, and C are more easily differentiated when grading a musical program, while grades D, E, and F are most significant with respect to speech. Hence, if a given program sample contains both speech and music, it is readily graded, whatever its quality may happen to be.

Signal-to-Noise Ratios

The signal-to-noise ratio corresponding to each program sample or listening period was obtained before the period from the radio-noise-meter readings, first with the program-modulated carrier on and the noise stopped and then with the noise on but with the carrier off. The ratio between these two readings is the signal-to-noise ratio, which will be expressed in decibels in this paper. It was assumed that both signal and noise remained at constant levels between the time of measurement and the listening period. The noise sources were so chosen as to justify this assumption with respect to them. The program level was sensibly constant, as is usual with the type of program ordinarily broadcast in the morning hours. Signal-to-noise ratios were determined for each program sample with each of the three radio-noise meters.

[4] C. M. Burrill, discussion of paper by C. V. Aggers, D. E. Foster, and C. S. Young, "Instruments and methods of measuring radio noise," *Elec. Eng.*, vol. 59, pp. 178–192; March, 1940.

The Tests

Each of thirty observers, most of them untrained in listening tests, graded the "quality of reception" of twenty-four listening periods, comprising four different signal-to-noise ratios with three different kinds of noise, reproduced on two different receivers. The various signal-to-noise-ratio values were presented in random order, and the observers were not informed of these values until after the completion of all the grading. Table I gives the order of the test periods, together with the corresponding signal-to-noise-ratio values as determined with each of the three radio-noise meters.

TABLE I

Test	Noise Source	Receiver Type	Noise-Meter Signal-to-Noise Ratio—Decibels		
			A	B	C
1A	Commutator	Console	28.9	28.5	33.2
1B	Commutator	Console	19.2	19.7	25.0
1C	Commutator	Console	45.8	45.4	50.1
1D	Commutator	Console	54.8	54.4	59.1
2A	Razor	Console	40.6	40.6	50.1
2B	Razor	Console	14.6	12.6	19.2
2C	Razor	Console	48.6	48.6	58.1
2D	Razor	Console	22.9	21.2	29.3
3A	Relay	Console	34.5	36.1	46.6
3B	Relay	Console	42.5	44.1	54.6
3C	Relay	Console	19.2	18.1	28.5
3D	Relay	Console	12.0	10.1	22.1
4A	Commutator	Table	28.9	28.5	33.2
4B	Commutator	Table	45.8	45.4	50.1
4C	Commutator	Table	54.8	54.4	59.1
4D	Commutator	Table	19.2	19.7	25.0
5A	Razor	Table	40.6	40.6	50.1
5B	Razor	Table	22.9	21.2	29.3
5C	Razor	Table	14.6	12.6	19.2
5D	Razor	Table	48.6	48.6	58.1
6A	Relay	Table	12.0	10.1	22.1
6B	Relay	Table	19.2	18.1	28.5
6C	Relay	Table	34.5	36.1	46.6
6D	Relay	Table	42.5	44.1	54.6

The Judgments of the Listeners

When comparisons were made of the grades given to reception with the two different receivers, for the same type of noise and the same signal-to-noise ratio, no signicant differences were noted. Therefore, the data for each pair of tests with the two receivers were combined to give a total of sixty grades for each test condition.

The results for one typical test condition are shown in the form of a bar diagram in Fig. 2. Such a diagram is one way of representing what the statistician calls a frequency distribution. It indicates that a group of entities, in the present case, listeners, has been classified with respect to some specified attribute, in the present case, the grade which each gave to a particular program sample, and gives the number of the entities falling in each class. This number per class is called the frequency, indicated more specifically here as "frequency of occurrence" to distinguish it from the frequency of an alternating current.

The frequency distribution represents a considerable condensation of data, since no indication is given as to which entity falls in which class. Ordinarily this represents no loss of *useful* information, and is a great convenience. Thus, in the present instance we are not

interested in the grades given by the individual listener, but rather in the grades of the group as a whole.

Statisticians have found that nearly all frequency distributions may be grouped into a few basic types. One of the simplest of these types is so often encountered that it has been called the normal frequency distribution. As we shall see, the data plotted in Fig. 2 may be adjusted to yield a distribution of this type. The data represented by a frequency distribution may be condensed further by calculating, in a manner depending on the type of distribution, certain parameters and using these to characterize or represent the distribution itself. Two kinds of such parameters, applicable to most types of frequency distribution, are those indicating "central tendency" and those indicating "dispersion." In the language of our present example, we may represent the data for each listening test by two quantities which answer the questions, what is the convergent trend of opinion, and how much do opinions differ?

It is easy to see that the opinions trend toward about D+ for the test to which Fig. 2 applies, but the statistician requires a more precise determination of the trend or central tendency. From the number of statistical quantities useful in indicating the central tendency of such data we shall choose the simplest and the one most used—the arithmetic mean. For this purpose we must assign, somewhat arbitrarily, numerical values to the classes or grades. The data can then be

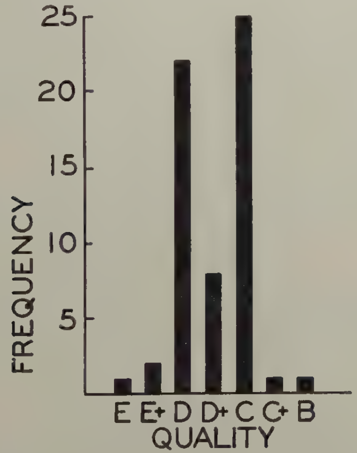

Fig. 2—The listeners' judgments for a typical test summarized in the form of a bar diagram. Noise meters A and B, commutator noise, signal-to-noise ratio = 28.7 decibels.

plotted to a continuous abscissa scale, as in Fig. 3, using the following equivalents:

F	F+	E	E+	D	D+	C	C+	B	B+	A
0.0	0.5	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0

The arithmetic mean of all the grades may now be computed and expressed on this scale. It is found, for the data represented by Fig. 3, to be 2.51 or very nearly D+ as was to be expected.

The most commonly used statistical measure of dispersion, that is, the scattering of the data, is the standard deviation, or the root-mean-square deviation from the arithmetic mean. This is computed by summing the squares of the deviations of each grade from the average grade and extracting the square root of the sum. The standard deviation of the data represented by Fig. 3 is 0.566 or about half a grade.

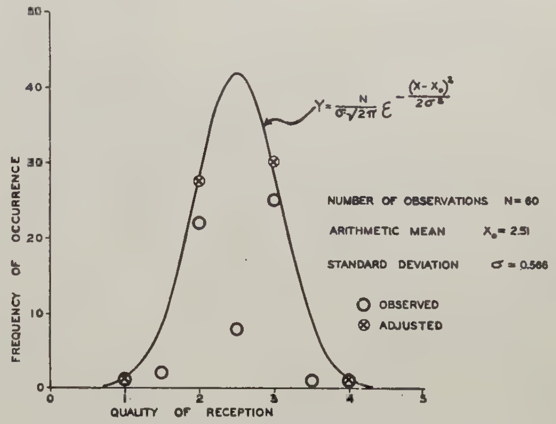

Fig. 3—The data of Fig. 2 expressed in the form of a frequency-distribution curve.

Suppose we fit a smooth curve to the points plotted on Fig. 3. Such a curve is called a frequency-distribution curve or simply a frequency curve. It must be noted that this is not at all what a radio engineer usually means by a frequency curve. Obviously many different curves could be drawn through the rather scattered points of Fig. 3. A freehand fitting would be a strain on the judgment. So we show in the figure a particular type of curve, called the normal-distribution curve, having the same arithmetic mean and standard deviation as we computed from our data. This curve corresponds to the equation

$$y = \frac{n}{\sqrt{2\pi}\cdot\sigma}\cdot\epsilon^{-(x-x_0)^2/2\sigma^2}$$

where x = quality of reception

y = frequency of occurrence

$n = 60$ = the total number of observations

$x_0 = 2.51$ = arithmetic mean of the observations

$\sigma = 0.566$ = the standard deviation of the observations

It is noted that this theoretical distribution curve fits the observational data fairly well except for the point at quality = 2.5. The fit is close if a plausible modification is made in the data. It is evident that the listeners tended to favor the even grades A, B, C, etc., over the intermediate grades B+, C+, etc. Out of a total of 720 grades for all the tests, less than 11 per cent were intermediate grades, whereas if the listeners were not discriminating against intermediate grades one would expect about 50 per cent. It we divide up the intermediate-grade judgments between the two adjoining even grades, in proportion to the numbers of judgments for those adjacent grades, we obtain the following modified frequency tabulation:

Grade	E	D	C	B
Modified Observed Frequency	1.09	27.66	30.21	1.04
Theoretical Frequency	1.23	28.2	29.0	1.33

TABLE II

Noise Source	Noise Meter	Signal-to-Noise Ratio, Decibels	Quality of Reception: Arithmetic Mean	Standard Deviation	Probable Error: Single Observation	Probable Error: Mean of 60 Observations	Correlation: Regression Coefficients a	b	Standard Error of Estimate	Coefficient of Linear Correlation
Direct-Current Commutator Motor	A & B	19.4	1.742	0.511	0.347	0.0445	0.154	0.0826	0.612	0.882
		28.7	2.508	0.566	0.385	0.0497				
		45.6	4.025	0.602	0.409	0.0524				
		54.6	4.592	0.725	0.493	0.0631				
Razor	A & B	13.6	0.842	0.566	0.382	0.0493	−0.259	0.090	0.648	0.889
		22.0	1.950	0.472	0.318	0.0411				
		40.6	3.167	0.667	0.450	0.0581				
		48.6	4.233	0.735	0.496	0.0640				
Relay	A & B	11.0	0.542	0.519	0.350	0.0452	−0.3599	0.09625	0.590	0.903
		18.6	1.683	0.533	0.360	0.0464				
		35.3	2.892	0.611	0.412	0.0532				
		43.3	3.858	0.608	0.410	0.0529				
Direct-Current Commutator Motor	C	25.0 33.2 50.1 59.1	Same as				−0.328	0.0847	0.611	0.881
Razor	C	19.2 29.3 50.1 58.1		for Meters			−0.615	0.0807	0.646	0.890
Relay	C	22.1 28.5 46.6 54.6			A and B		−1.305	0.0935	0.508	0.929
Averages				0.593	0.401	0.0516	—	0.08796	00.6025	0.896

The closeness of this fit of our data to the theoretical normal-distribution curve permits and encourages us to make certain assumptions from which useful conclusions can be drawn. Thus we postulate a statistical population of all or a very large number of people like those who participated as listeners in our tests, and we suppose our listeners to comprise a fair sample selected at random from this population. We further assume that if the judgments of this entire population regarding a given radio-program sample could be tabulated they would turn out to be normally distributed, that is, distributed in accordance with the normal-distribution curve. It is also assumed that the arithmetic mean of the judgments of our sample group of listeners would coincide with the arithmetic mean of the judgments of the entire population.

Now, under these assumptions, we can estimate from our data the standard deviation of the entire population, in accordance with the formula

$$s \cong \sqrt{\frac{n}{n-1}} \cdot \sigma = \sqrt{\frac{60}{59}} \times 0.566 = 0.571.$$

From this value the probable error of a single judgment, and the probable error of the mean of a sample of n judgments, in each case selected at random from the population, can be calculated. Thus

probable error of a single observation $= 0.6745s = 0.6745\sqrt{\dfrac{n}{n-1}} \cdot \sigma = 0.385.$

probable error of mean of n observations $= 0.6745\dfrac{s}{\sqrt{n}} = 0.6745\dfrac{\sigma}{\sqrt{n-1}} = 0.0497.$

The interpretation of these values is as follows:

(1) The odds are even (probability 0.5) that the judgment of a single listener chosen at random, regarding the particular program sample we have been considering, would fall within ± 0.385 of a grade of the arithmetic mean obtained in our test.

(2) The odds are even that the arithmetic mean of the sixty judgments obtained from a second similar group of listeners chosen at random would fall within ± 0.0497 of a grade of the arithmetic mean obtained in our test.

In Table II are given the results of the grading for each of the four noise levels and for each of the three types of noise. The arithmetic mean of the grades, the probable error of this mean, and the probable error of a single grade are tabulated. It is indicated that very satisfactory reliance can be placed on these data.

The Correlation of the Radio-Noise-Meter Indications

In referring to Table I it is noted that the indications of radio-noise meters A and B were very nearly alike. Therefore the readings of these two instruments were averaged for use in the subsequent analysis.

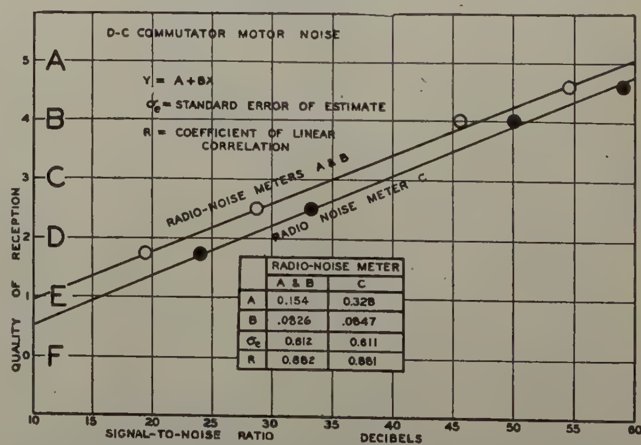

Fig. 4—Quality of reception as a function of signal-to-noise ratio, for commutator-motor noise and for two types of radio-noise meter.

Fig. 4 shows a plot of the quality of reception determined by the arithmetic mean of the listeners' judgments, as a function of the signal-to-noise ratio in decibels indicated by the radio-noise meters. It represents the tests with direct-current commutator-motor noise; similar plots for the other tests are shown in Figs. 5 and 6. It is evident that the observed points shown in Fig. 4 can be represented very well by two straight lines, one for each type of radio-noise meter. Usually such straight lines would be drawn in simply by inspection. However, in the present instance we wish to determine a numerical measure of how well these lines represent the data, and so we must be careful to obtain the "best" fit. This is obtained by the method of least squares, by which the sum of the squares of the deviations of the observational points from the line is made a minimum. The lines shown in the figure, which were obtained in this way, would be called by statisticians linear regression lines; the coefficients a and b in their algebraic expression, the regression equation $y=a+bx$ would be called regression coefficients.

The usual measure of how well a regression line fits the data is the "standard error of prediction." It is equal to the square root of the sum of the squares of the deviations of all the values of the dependent variable—"quality of reception" in the present case—as predicted by the regression equation, from the corresponding observed values. This quantity σ_e which is made a minimum by the use of the method of least squares in determining the regression equation, is a measure of the error to be expected in estimating a value of y from the corresponding value of x and the regression equation.

Without any knowledge of x, the best estimate one could make of a particular value of y would be the arithmetic mean of all the y values, and the error to be expected would be measured by the square root of the sum of the squares of the deviations of all of the y values from this mean. This latter quantity is called

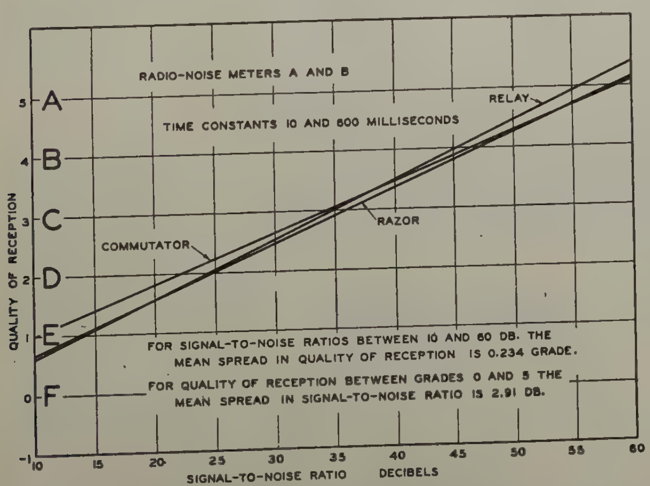

Fig. 5—Quality of reception as a function of signal-to-noise ratio as read by radio-noise meters A and B, for three different kinds of noise.

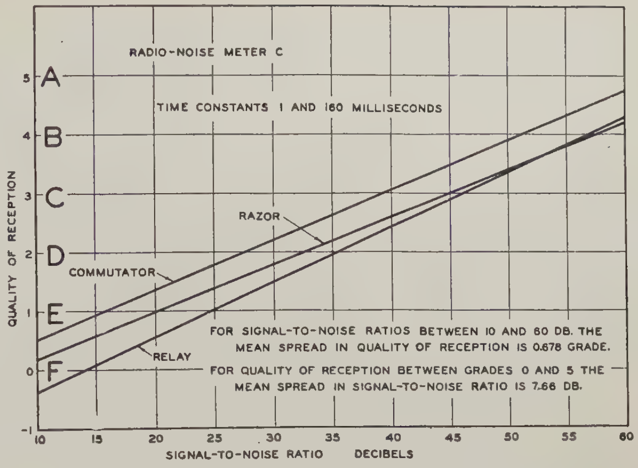

Fig. 6—Quality of reception as a function of signal-to-noise ratio as read by radio-noise meter C, for three different kinds of noise.

the standard deviation of the y's or σ_y. Then the ratio σ_e/σ_y indicates how much a knowledge of the corresponding value of x reduces the error to be expected in predicting a particular value of y. The error is reduced only if x and y are correlated. The usual measure of this correlation is the coefficient of linear correlation r, given by

$$r = \sqrt{1 - (\sigma_e/\sigma_y)^2}.$$

The values of the regression coefficients, the standard errors of estimate, and the coefficients of linear correlation, for the case chosen for illustration—that of direct-current commutator-motor noise—are exhibited in Fig. 4, along with plots of the mean judgments and the regression equations. The values of these parameters for all the cases are listed in Table II.

Comparison of the Radio-Noise Meters

The two types of radio-noise meter may be compared in two ways using the foregoing data. First, it is observed by comparing the standard errors of estimate given in Table II that the data for radio-noise meter C are fitted by linear regression curves a trifle more closely than are the data for radio-noise meters A and B. Second, the regression lines pertaining to radio-noise meters A and B are more nearly the same for all three types of noise than are those for radio-noise meter C. This is noted by comparing the values of signal-to-noise ratio corresponding to the different grades of quality of reception calculated from the regression equations. These values are given in Table III, and indicate an average range of variation for the different types of noise of 3.2 decibels for radio-noise meters A and B and 7.8 decibels for radio-noise meter C.

The comparison may be made graphically by means of Figs. 5 and 6, which show, respectively, the three regression lines for radio-noise meters A and B and for radio-noise meter C. These figures give what are in effect radio-noise-meter calibration curves, there being a separate curve for each type of noise meter

and kind of noise. For an ideal noise meter, one curve should suffice for all types of noise. The average separation of the curves for radio-noise meters A and B within the range covered by the tests, as shown in Fig. 5, is 2.9 decibels[5] in signal-to-noise ratio, or 0.23 grade[6] in quality of reception. The corresponding values for radio-noise meter C, as shown in Fig. 6, are 7.7 decibels[5] or 0.68 grade.[6] These values may be called the noise-meter correlation errors, since they are indications of the degree of failure of the instrument readings to indicate quality of reception independently of the kind of noise.

TABLE III

Quality of Reception		Signal-to-Noise Ratio in Decibels							
		Noise Meters A and B				Noise Meter C			
Grade	Numerical Scale	Commutator	Razor	Relay	Range	Commutator	Razor	Relay	Range
A	5	58.6	58.5	55.9	2.7	62.9	69.6	67.5	6.7
B	4	46.6	47.4	45.4	2.0	51.1	57.2	56.8	6.1
C	3	34.4	36.2	35.0	1.8	39.3	44.8	46.1	6.8
D	2	22.3	25.1	24.6	2.8	27.5	32.4	35.4	7.9
E	1	10.2	14.0	14.2	4.0	15.7	20.0	24.6	8.9
F	0	−1.8	2.9	3.8	5.6	3.9	7.6	14.0	10.1
Average Range Decibels					3.2				7.8

Significance of the Observed Correlation Errors

It would be difficult for a listener to detect a 2.9-decibel change in level of a broadcast program, unless the change were made very suddenly. It is reasonable to expect that a 2.9-decibel change in signal-to-noise ratio would be at least as hard to detect.

Since the fundamental determination of quality of reception is by the judgment of listeners, the noise-meter correlation errors should be compared with the errors to be expected in the basic subjective, or listening method. Now the average, for the twelve test conditions, of the standard deviation of the listeners' judgments, was 0.593 grade (see Table II). This corresponds to a probable error, for the judgment of a single listener, of ± 0.40 grade, or a spread of 0.80 grade. Since this value is more than three times the correlation error of noise meters A and B, it may be said with assurance that these instruments are more reliable than the judgment of a single listener. In fact, it may be calculated on the same basis that radio-noise meters A and B are equivalent in accuracy to a jury of twelve listeners. On the same basis, noise meter C is a little more accurate than the judgment of a single listener.

[5] These values were obtained by finding the area bounded by the extreme calibration curves of the group and abscissas corresponding to quality grades A and F (0.0 to 5.0). They are therefore slightly different from the similar values given in Table III, which were obtained by averaging the separations at the five even grades A, B, C, D, and E.

[6] Obtained by finding the area bounded by the extreme calibration curves of the group and the ordinates corresponding to signal-to-noise ratios of 10 and 60 decibels.

Conclusions

The scale of grades A, B, C, etc., used for the subjective determination of quality of reception is shown to be very satisfactory from the statistical point of view.

Radio-noise meters of either of the two types tested may be used to indicate quality of reception on this scale with reasonable accuracy, with linear calibration curves (regression lines) which do not vary greatly for different types of noise. Radio-noise meters A and B, embodying the time constants adopted by Joint Co-ordination Committee, had somewhat smaller correlation errors than radio-noise meter C employing the modification of these time constants under consideration in Canada. All three of these instruments are believed to give indications proportional to quality of reception much more uniformly for different noise waveforms than any instrument previously available, although comprehensive tests corresponding to those here reported have not been carried out for other instruments.

Finally, the adequacy of the Joint Co-ordination Committee specification is indicated by the satisfactory uniformity of performance, in these tests, of radio-noise meters A and B, built to conform to it by different manufacturers. The validity of this conclusion when noise waveforms of the types used in the present tests are concerned has not been challenged, but subsequent investigations have indicated that the Joint Co-ordination Committee specification is not entirely adequate when noise waveforms consisting of widely separated sharp peaks are involved. In the latter case the transient characteristics of the automatic-volume-control circuit play a dominant role in determining the instrument response, and it therefore appears that these characteristics should be more definitely specified. This matter has been further discussed by the author elsewhere.[7]

Acknowledgments

The author expresses his thanks and appreciation to all those who assisted in the tests and to those who co-operated in the writing of the paper; especially to L. C. F. Horle for his interest and enthusiasm in sponsoring the tests, to D. E. Foster for planning and carrying them out, and to W. C. Morrison, who checked the computations.

[7] C. M. Burrill, "Some observations concerning the transient behavior of radio-noise meters." Presented, Rochester Fall Meeting, Rochester, N. Y., November 12, 1941.

A Symposium On

RADIO IN THE WAR EFFORT

Rarely does the radio field, speaking as an integrated community, clearly and inspiringly express its aims and accomplishments. At the June, 1942, Convention of the Institute of Radio Engineers at Cleveland, Ohio, our membership listened as the responsible leaders of opinion in the major divisions of radio contributed to a constructive and unified discussion. The Presidents of our engineering fraternity, The Institute of Radio Engineers; of the grouped constructors of equipment, the Radio Manufacturers Association; and of the affiliated broadcasters, the National Association of Broadcasters; conclusively proved on that occasion that, beneath the dust and turmoil of our daily activities and relationships, rests a solid foundation of common interests and mutual respect. It is fortunate that their thoughts were expressed, since the words of such speakers may well lead to closer understanding and greater co-operative effort between all portions of the radio field in the years to come.

The Editor

War Contributions of Radio Manufacturing*

PAUL V. GALVIN†, NONMEMBER, I.R.E.

I AM indeed delighted in having this opportunity to be with you this afternoon and participate in this Panel covering radio's efforts in the war.

From my observation of those on the Panel, the broadcast phase of radio is very well represented. Therefore, I shall confine my remarks to the manufacturing and apparatus phase. In my reference to engineers, I refer particularly to design and development engineers. In passing, I only wish to state in regard to broadcast that the manufacturers are very conscious of their responsibility in keeping the radio sets in the hands of the American public in a satisfactory state of repair and in operation. We are constantly concerned with the matter of maintenance and replacement parts. Our Association maintains an active committee on this subject. We are even investigating the future prospects of utilizing Boy Scouts for the necessary servicing of the radio sets in the hands of the public, when the service organization as we know it has completely gone to war.

THE THREE PRESIDENTS

Neville Miller, National Association of Broadcasters; Arthur Van Dyck, Institute of Radio Engineers, and Paul V. Galvin, Radio Manufacturers Association.

The radio manufacturing industry was in the vanguard of those industries who recognized their responsibility in the Defense Program as launched in the summer of 1940 by President Roosevelt. We took this as the initial opening phase and chartered a course of industry in the preparation for war. Immediately following the launching of the Defense Program in 1940, discussions took place in group meetings of the Radio Manufacturers Association searching for the most intelligent approach for industry guidance in the defense effort. Up to this time the radio industry—like all America—was primarily concerned with their peacetime pursuits. A few large companies were making some radio apparatus for war purposes.

* Decimal classification: R560×R700. Original manuscript received by the Institute, June 29, 1942.

† President, Radio Manufacturers Association, Washington, D. C.

Limited allocation of funds to our Army and Navy for the purpose of radio equipment development severely limited the activity of the Army and Navy on radio engineering and production prior to the summer of 1940. With the invasion of the lowlands in France in May and June of 1940, the importance of radio and the part it was to play in modern warfare became very apparent. The precise co-ordination of all of the service arms of the German Army was accomplished by the effective use of radio.

When wartime radio problems hit the industry in the summer of 1940, we found them in a highly amorphous state. The engineering, manufacturing, and procurement of radio apparatus needed further prescription of the broad problems we were to tackle.

In World War I, we had no radio industry as we know it today. This is the first entry of radio as an industry in a war. True, radio was used in World War I but not in the sense nor to the extent we know it today. The Navy did depend on wireless but aside from that, radio was a minor factor in the last war. Today radio is used as communication by all of the service arms. Today each tank contains an average of $5000 worth of radio equipment. A single heavy bomber carries on the average of some $50,000 worth of the most complete and delicate devices. The protection of ships, troops, and even our homeland is now very much involved in the use of radio. Radio also enters very materially into the all important "war of nerves." The backbone of enemy spy activity is radio. Radio also plays a very important part in maintenance of public morale.

By September, 1940, it became apparent to the radio industry that it had a big task to perform in the Defense Program. The magnitude of the job was emphasized by the projection of a radio apparatus program amounting in excess of one billion dollars. Up to this time a half-dozen companies were making the few millions of dollars worth of apparatus the Army and Navy bought.

Gradually the problems began to spread through the industry. It was impossible to show results over night. People had to become adjusted. Organizations had to become adjusted. Engineering departments of the radio industry had to become adjusted. Research and procurement divisions of the Army and Navy expanded. The engineering activity as a whole from a military standpoint had to take on a new and increased tempo. The efforts of management of the radio industry started channeling into its new defense grooves.

While this was going on during 1941 civilian production continued—and some, unfamiliar with the whole problem, advocated disruption of the civilian production organizations. But the industry leaders as well as the Army and Navy and War Production Board officials knew that the industry production organizations must be kept intact as going concerns until the engineering and procurement problems of this whole, new vast program shook down to a state where they were in shape to go into production on a scale to utilize the production facilities of the industry. A vindication of this policy was contained in the recent report of the Senate investigation of the WPB. This Committee investigated the activities and reported on the industry. At the conclusion of the report the Senate Committee made a statement to the effect that in view of the facts presented, criticisms in regard to the radio industry were not well founded.

That war is destructive is axiomatic. The purpose of war is the destruction of enough of the assets, human and material, of the enemy to force him to give up the struggle. But war destroys not only lives and property. War causes a great alteration of ideas and habits—peacetime ideas and habits change to wartime ideas and habits. For the very best war effort we must become war minded. This means, not just idly thinking of the war or wondering who's going to win and when the war will end. We must marshal real fighting ideas and turn our works and our habits of working into the groove that will best aid in our country winning the war. Those who have not negotiated this mental bottleneck as yet haven't made the proper adjustment to where they can make their very best war effort.

Engineers too often are inclined to view their job narrowly. The engineer too often prefers to resolve his problem within well-defined limits and within a field whose structure is clearly defined. The engineer is too prone to want plenty of time to conclude a project with finality in the prescribed fashion—that's out, positively, for the duration of this war. Your problems will never be finished. You will be called upon constantly to explore new horizons in the radio and electronics field. You will be called upon for better and quicker answers. Courageous enterprise on the part of the engineer will be a major contribution to our winning the war. I might go so far as to say that without courageous enterprise on the part of our engineers, we might lose the war. We must have new and better models of everything, planes, tanks, boats, guns, and radio equipment. If we don't, our smart enemy will have better equipment and apparatus and we will be outclassed.

You are being called upon and will continue to be called upon throughout this entire war to strain to the breaking point for the deliverance of energy, effort, and brains, the like of which you have never done before. You may have thought you were very busy many times before, but this war effort calls for a personal and social readjustment and sacrifice of all engineers to where they must give complete devotion tenaciously to their task. The engineers are being relied upon to come up with intelligent answers which the manufacturers can put into production in a hurry, and thus put into the hands of the Army and Navy large quantities of material for greater and more effective striking power.

In the war effort, radio engineers have to be ready and willing at all times to tackle the problem of substitution of materials when it comes up, regardless of how annoying it may be. It is part of the game. One way to keep clear of the critical material problem as best you can is to design away from it. And in your design activity I caution you to watch this critical material problem very carefully. It will save you, your manufacturer, and the Army and Navy many headaches. The mechanical engineer and the electrical engineer in this effort, more than ever before, must have a greater spirit of co-operation, one unto the other, in quickly working out design problems. The design engineer should be more conscious of the manufacturing and tooling problems. These can be serious bottlenecks in getting apparatus through the plant in quantities, and on time.

Today the radio development and design engineer has to be more cognizant of the other fellow's problems than ever before. The effectiveness of our industry war effort depends upon the proper integration of all of our efforts—and these efforts can be so much more fruitful all along the line when broad thinking is applied to the development and design of the apparatus when it is still in the hands of the engineer.

Radio men are up against some clever engineers in the radio and electronics field in both Germany and Japan. An examination of the technical literature will show you that—and the Nazis have turned out apparatus which will command your attention and challenge. I wonder sometimes, if you men thoroughly realize the importance radio is destined to play in the winning of this war. The whole pattern of war tactics and strategy has been altered by the use of radio communication and radio direction finders. The co-ordination of land, air, and sea forces is accomplished by radio. Protection from the enemy and firing accuracy is accomplished by Radar. It has been said that in the aerial battle for Britain in the fall of 1940 radio direction finding apparatus which we, in this country, call Radar, was a prime contributing factor of the Royal Air Force's maintaining superiority in the air over the Nazis with a much smaller aggregation of flying equipment. You are alive, I am sure, to your war effort responsibilities but I implore you to do more. You must do more. We all must do more if we are to win this war.

The management group as a whole in whose hands the war production effort of this radio industry has been entrusted, are fully conscious of their very serious responsibility in this program. They have stripped their plants for necessary action and are producing apparatus in huge quantities. They realize they will be continuously pressed to do more and better. They are just now feeling the acceleration from their early efforts. They are prepared and will meet the requirements and beat schedules. I am fully confident the radio industry will come through for the Army and Navy on every score. It is a big order, I know, when we realize the magnitude of this vast radio and Radar program. But the radio manufacturers are used to "licking" big problems; they know their problems in this war effort and they will be solved. To you radio engineers who are "in the groove" and making your grand contribution to this great effort toward our winning the war, "Hat's off to you and keep up the good work"—and that, I am glad to say, goes for most of you. To you few who are not yet "in the groove," giving your very best and your all in this war effort—I say, "break that old mental bottleneck—dust off the cobwebs and get in there with some good intelligent licks." Your brainstorm may be the "rabbit out of the hat" that will make a most valuable contribution to this effort.

Industry, by its deeds in the war effort, is standing the business "baiter" back on his heels. The critics of reputable business seem to have had a "field day" before the war. Today their demagoguery is being answered by action. The production job being done by industry in this war effort is a vindication of the private-enterprise system. The public is, and will continue to be, very much impressed with the job industry is doing. These accomplishments of assembly, process, and method are all basically engineering. Let's be sure when these accomplishments are recorded in history, that the radio industry can proudly look back on its record.

Yesterday morning I received a communication from James S. Knowlson, Director of the Division of Industry Operations of the War Production Board in Washington, D. C. In that communication the following paragraph is of specific interest to the radio engineers:

"Of course the radio industry has a tremendous job ahead and probably a good deal of grief because the art changes so rapidly it is hard to keep up with the requirements. Certainly, if necessity is the mother of invention, we are going to see a lot of new things in the radio and Radar developments, and I imagine that when we go back to television we are going to find that most of the standards that have been made are obsolete. In the meantime, war production is the big thing, and as you say, it looks like quite a job."

Work hard during the war; your fun is coming after the war is over. With all the new materials, new tubes, and new ideas developed during the war you are going to have a picnic shaping them into playthings for commercial and civilian application. There will be no "*status quo ante bellum*" for the radio engineer.

What Radio Broadcasting Means in the War Effort*

NEVILLE MILLER†, NONMEMBER, I.R.E.

AMERICA is not a warlike nation. Our natural impulses are humane and charitable. We have the inborn determination which is necessary to make good warriors when forced to fight but with it all have a spirit of fairness which was evidenced by our treatment of the eight saboteurs recently caught red handed who were planning to kill and destroy. I mentioned this because what radio means today in the war effort must be interpreted in the light of our background—must be interpreted as radio developed during times of peace, and not developed to be used as a war instrument. Let us look back briefly at the twenty years between the two wars, years of disarmament conferences when men were trying to get away from war. The Army and Navy undoubtedly were making such plans as they could against obstacles. Our gold was moved to Fort Knox amid the cheers and jeers of our comedians.

I remember visiting Fort Knox in company with a Congressional Committee in 1936 and seeing, at that time, the entire tank force of our Army. Considering the importance which tanks now play in warfare, it is hard to realize that even six years ago the Army had difficulty in securing funds for experimental tanks from a reluctant Congress which was reflecting the wishes of a reluctant people.

What is the history of radio? We were granted the right by the government to use a few wavelengths. There was no government money provided for development; no appropriations, such as road appropriations or harbor appropriations; no subsidies by mail contracts and no second-class postal privileges. All that we asked was to be let alone and during the last twenty years, thanks to your engineering ability and to the salesmen and program builders and to Mr. Galvin's associates and Mr. Jett's associates we built highways into the homes of the American people. Today there are 56,000,000 sets in 85 per cent of the homes and the American people listen to the radio.

* Decimal classification: R560. Original manuscript received by the Institute, August 24, 1942.
† President, National Association of Broadcasters, Washington, D. C.

Then war came and overnight radio was able to deliver to the government without cost the greatest means of mass communication the world has ever seen. No tooling-up period was needed, no appropriation bills were passed through Congress. It was all done so rapidly and so quietly that it was taken for granted.

There are no four-minute speakers today as there were in the last war. There are no trains taking groups around the country to campaign. All that today is being done by radio. There are just as many campaigns—maybe more—campaigns of recruiting, for the sale or war bonds, salvage, rubber, and all the others with which you are so familiar. But the burden of all these campaigns is being borne cheerfully and successfully by radio.

We face many problems in a difficult age but our greatest problem is not engineering, not the procurement of equipment, not the securing of business, not Mr. Petrillo—our greatest problem is keeping our social developments abreast of our engineering development. I have every confidence that the engineering skill of America will solve all of our engineering problems, but we must realize that unless our social development is kept abreast of the engineering development, the net result may be disastrous.

The history of America is the story of a great nation. However, we have been extravagant. We have wasted our forests, our public lands, our oil. Today we have received an asset in radio equal to all of our natural assets. You and I were not parties to what has happened in the past but you and I are going to have a part in writing the chapter of the future. Let us avoid the past mistakes; let us conserve our assets. Let us see that radio is used to make America's future a brilliant future. It is a grand chapter and let us be proud of the way we write it.

Radio Engineering in the War Effort*

ARTHUR VAN DYCK†, FELLOW, I.R.E.

MOST analysts agree that the most important element in this war is aviation, that next most important is ordnance, and in third place is radio. Each of these elements has shared in the general and rapid advance in technology of recent years. As a result each has developed rapidly in capability, complexity, and degree of effect on the practice of war. Just how rapid and how great that development has been is clearly seen by comparing their present capabilities with those of World War I. Aviation and radio are similar in that both had their introduction to war

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† President, Institute of Radio Engineers, New York, N. Y.

in World War I, and then revealed clearly their enormous potentialities.

We are concerned here with radio only, and will find it interesting and helpful to compare its uses in World War I with those of today. To those now under forty years of age, and who, therefore, were under fifteen years of age when we entered World War I, the radio conditions of that time are unknown by personal experience, and must seem antediluvian. There was no broadcasting, no transmission of pictures by radio, and transoceanic radiotelegraphy was sporadic and unreliable. Practical radiotelephony had just been born. Even radiotelegraphy was only five years old, in the sense that only for that length of time had passenger vessels been required to have wireless and operators to be licensed.

What a contrast is presented by today's conditions! Now transoceanic radiotelegraphy and telephony both are giving every-day reliable service, and broadcasting covers the earth. Television and entry into the higher-frequency spectrum are repeating the conditions of new possibilities in war which were brought to the last war by the vacuum tube and radiotelephony. New and great significance results from the application of space radio principles to uses other than communication, to which radio was confined previously, such as the now well-known radio plane detectors. The work of the radio engineer would have inestimable value if radio were used only for communication purposes in ships, planes, and tanks. With other uses added, it takes on importance second only to that of the planes and guns themselves. It seems to me correct to say that never before in the history of science, has any branch had as varied, ramified, and powerful utilization as radio does today. The radio engineer of today has a right to stick out his chest and feel important, even as he has the obligation to feel and realize his responsibility to use to the fullest advantage his opportunity for unusual service in this critical time.

As this country approached war in 1941, the radio industry encountered its first war problem in the shortage of certain critical materials upon which radio apparatus had depended from its beginning. Radio engineers met that problem so successfully that in spite of shortage and substitution, more receivers were manufactured in 1941 than in any previous year.

Then came conversion from civilian home-entertainment receiver production to the manufacture of radio apparatus for war. This was really difficult, because military radio involves degrees of complexity, precision, and ruggedness not known in civilian radio. It meant changing engineering and factory practice from designs utilizing crude tolerances to ones of high precision, both mechanically and electrically. It meant changing from apparatus required to meet only the range of conditions between parlor and kitchen to apparatus capable of working reliably in the stratosphere and in the equatorial desert. The burden of this change fell most heavily upon the radio engineers. And well have they carried it—without adequate sleep, without furloughs, without rewards, or medals, they are doing the job.

I have not recited this story just to record what has been done so far, or to pat the profession on the back. I have done so in order to silhouette what remains to be done, because if we are to continue on to the greater tasks ahead, with maximum efficiency, we must realize where we fit in the picture, and how basically vital radio is in this war. Furthermore, to advance and improve we do not need to dwell upon the good things we have done or are certain to do, but instead we need to know the bad, and to correct them.

The new mobility of attack, on land, on sea, and in the air, is possible only by use of radio communication. The companion mobility of successful defense is dependent likewise upon radio communication. In World War I, radio communication was an adjunct; in this war it is a vital necessity. In addition, we have now the applications of radio techniques to new instruments and weapons, thereby broadening the field of radio to limits not yet clearly seen. How can realization of this situation assist toward better execution of the tasks ahead?

The task ahead is not merely the invention and development of new devices, but even more important, is wise utilization of the ones we already have. Utilization of the things we have is not in the hands of engineers alone; in fact, it is mostly in the hands of others. Therefore, it devolves upon engineers to educate those others in the facts of technical life as rapidly and forcefully as possible. In peacetime we can allow the time needed for sufficient education to percolate slowly through the minds of others involved, with any amount of accompanying confusion and loss. In war we cannot with safety allow that time or that confusion and loss.

The advances in radio are but a part of the advance of technology on many fronts. In aviation, metallurgy, chemistry, and plastics, to name only a few, the advance has been so rapid that, as in radio, the nontechnical person could not understand the implications of the early stages until the final stages had been reached. So, for example, we have had various phases of aviation unappreciated until very recently, although aviation experts have understood them for many years. Unfortunately for the world, this condition did not exist in Germany. There, in Dr. Haushofer's incredible Institute, and in the German Army, was full realization of the new importance of technology and a thoroughly integrated utilization of it.

There are some examples of the kind of thing which we in radio need to do in our own industry, to improve our integration, co-ordination, and utilization of technical possibilities, and which can be mentioned in public forum. One is the utilization of the broadcasting system for public information, instruction, and control

in the event of air attack. So far this vast system has not been arranged for such use at all, in spite of its obvious availability and effectiveness. Let us examine this particular situation from the integrated technical viewpoint, which of course includes radio, military, and psychological factors, and see what conclusion would be reached.

Preparation of protection against air attack on this country has the following factors:

1. Attack will come unexpectedly.
2. Attack is almost certain to include new elements of attack (because this is past practice of the enemy in new attacks).
3. Rapid distribution of information to organized workers and the public regarding new elements of attack can mean the difference between catastrophe and successful resistance.
4. Instant calling or signaling to certain vital organized workers (such as auxiliary firemen and home guards) is necessary if they are to be of service quickly, as they must be to be effective.
5. The American people are not like Europeans with respect to news information. When anything happens they are strongly in the habit of finding out what it is —but quickly! They will not sit in ignorance. They insist upon being informed.
6. The American people are trained, as are no other people, to look to radio for quick news.
7. The American broadcasting system gives good communication into every spot in the country where people are located in any number.
8. Broadcast stations, in numbers sufficient to give adequate news dissemination, can stay on the air without giving aid to the enemy, if certain technical provisions are made, and if proper judgment is used in the information broadcast.

If these factors are correct, and every thoroughly experienced radio engineer knows that they are, why is not the broadcast system already set up and operating as a public war communications system, now that the whole public is in the front line of war? Simply because we lack integration of knowledge and decision. We engineers have not educated others to know that we have a system available which can be used without any objection. The public was not in the front lines in the last war, and did not need quick communication. This war is different, but we have not educated old-line thinking to know that now there is advantage in radio communication to the public comparable with its advantage in communication to tanks and planes. I hope that we can find a way to convince and coordinate the various agencies involved in the use of broadcasting before there is need to use it.

In the design and manufacture of war radio apparatus we will have plenty of difficult problems. We should try to the utmost to design and build with the maximum of standardization and the maximum of reliability. The problem of maintenance of military equipment will be one of huge and increasing difficulty. The enormous quantities of apparatus, the extreme complexities of many types, the scarcity of well-trained servicing personnel, all add up to a tremendous difficulty and even military failure of operations if the apparatus is not well designed and well built

The story of this war is being written into history. Some of it is behind us, but much more is ahead. The part which radio is playing and will play is being determined by the radio engineers. It is a vital part, and perhaps when the whole story is written it will be seen to have been a determining factor in the outcome. However it may be, we will be in there, doing our best to give all the assistance we can toward the destruction of the most barbarous organizations the world has ever known, in order that we may again proceed with peaceful, orderly development of things for the betterment of mankind.

Correspondence

The following letter merits the close attention of the readers of the PROCEEDINGS. None will question the writer's conclusion that technology can become the handmaiden of destruction in states dominated by ambitious, impulsive, and ruthless dictators animated by an unbridled passion for self-aggrandizement. Our readers are doubtless divided into camps on the question whether the engineer should, and could, guide or control the mode of utilization of his toil for the world's weal or woe. The views of our readers are invited, for their mutual information and benefit and as a possible broad guide to those formulating the policies of the Institute of Radio Engineers.

The Editor

TO THE EDITOR:

If Mr. Van Dyck[1] manages to arouse even a minority of the readers of the PROCEEDINGS to a sense of responsibility in the disposal and utilization of the products of their ingenuity, he will more than have justified his tenure of the office of president, provided, the minority be sufficiently roused to take the initiative in formulating some corrective measure.

The problem however is more fundamental than would appear from Mr. Van Dyck's article. The engineer has traditionally been the servant of mankind and has cheerfully taken his orders from anyone with sufficient driving force to assume the lead, and has never demanded of his overlords any too deep an appreciation of matters technical.

This state of affairs is, I submit, inherent in the nature of man. To be a successful technician a man must be capable—indeed he must enjoy the process—of subjugating himself to his problem, rendering himself receptive to all the evidence it presents, and applying the train of logic to his findings with complete disregard of his own personal desires as to the outcome.

Your leader, director, autocrat, exploiter, dictator (call him what you will) is fundamentally the opposite of this in character. For him to succeed his ego must always dominate. He deals not with physical properties but with the shifting intangibilities of men themselves, and, operating outside of logic, deals in their emotions—fear, pride, ambition and the like. This field may yet yield to scientific method, but with mankind as it is today, the technician naturally becomes the economic servant of the director type.

To advocate the delivery of some technical sense to our leaders is hardly a solution, yet some solution must be found if our children's children are to reap where we sow. Undoubtedly the forces that are now bearing down upon us will yield by burdensome trial and painful error the rare combination in our eventual leaders of initiative and driving force coupled with scientific integrity and technological insight. But to hasten that happy day, I would advocate the deliberate cultivation of a wider range of interest in every practicing engineer. If the majority of technicians were politically and economically well informed a body of opinion would arise whose influence could be a controlling one. If this opinion were scientifically informed it would of necessity be disinterested and act wholly for the good of the race. As initial moves toward such a condition the following might be contemplated.

(1) A campaign of education by this and all such technical societies to encourage members to interest themselves in the economic, political, artistic, and humanitarian aspects of their professional activities. To this end a special form of membership might be established to grant appropriate standing to those who, in addition to their technical abilities, have demonstrated their acquaintance with the relationship between their profession and the society in which it operates.

(2) The founding of a society of technicians to study the impingement of technical research on all phases of life, and to give the widest possible publicity to its findings among supporting technical societies. Too often the technician proceeds unperturbed along his chosen line of endeavor with no means of knowing to what final result his success may lead society.

(3) To organize a group to keep the technical man impartially advised of fundamental facts in current affairs having immediate bearing on his activities. Most of our opinions (which we flatter ourselves are as logical as our treatment of technical matters) are mere prejudices grown out of our emotional reactions to propaganda by one or other interested groups.

The emphasis in these somewhat tentative suggestions is not on any hope that change can come from the top but rather that it must originate with rank-and-file engineers, and this can only come about by their wider education in responsibility outside the present limits of their technical activities.

My salute to Mr. Van Dyck for his initiation of what I, for one, hope will prove a fundamentally important Institute activity.

L. T. BIRD
Vice Chairman
Montreal Section
Institute of Radio Engineers

[6] PROC. I.R.E., vol. 30, pp. 305–309; July, 1942.

Institute News and Radio Notes

TO BE OR NOT TO BE—IN UNIFORM

Bill paid a visit the other day to the laboratory where he used to work, on leave from his post deep in the heart of Florida. He had two of those beautiful Navy gold stripes on each sleeve and one of those new pre-Pearl Harbor ribbons on his chest—he was something! He carried an aura of valor, romance, and patriotism which seemed to spell "slacker" to every civilian suit of clothes in the lab. Had he been a recruiting officer, he could have signed up the whole staff right then and there. If you are a civilian in a laboratory today, you know what I mean.

But after he left I got to wondering about what would happen to the lab, and to the Navy, if the staff did sign up, or if even but a few more members like Bill, did so. That lab is developing and engineering the production of radio equipment which the Navy uses. And the Army too. The engineers who work there are a pretty special breed—years of training, years of experience in factory, field, and laboratory, and a teamwork which is knocking out results fast. In fact, there are fewer such fellows in the whole country than there are two stripers in the Navy.

The Navy and the Army cannot afford to have many more men leave the laboratories, if they are to keep on getting improved apparatus and new instruments until the war is won. But the men who have stuck to their jobs so far are getting restless under the lure of more direct service appeal, of uniforms and ribbons, and honors and preferments.

So, as an old timer who has been through it once, I want to appeal to the civilian radio engineers who are engaged in really vital war work, and who feel the pull of the uniformed services, to remember that their present work is vital, that they cannot be replaced by others in less than several years, and to sacrifice those honors and preferments which come readily to the uniformed services and not so readily to civilians. Such sacrifice will be worth while if it helps to assure the safety of freedom in the world, as it will. The balance is too critical right now to take any chances with our total effort.

The Army and Navy will need more men, and undoubtedly some more engineers from civilian ranks can be spared, but each case should be weighed carefully to determine whether the national effort will be better served by transfer or by remaining on the more prosaic civilian job. Certainly it is not to the national interest to have the wholesale desertion from laboratory ranks which seems to be looming.

Arthur Van Dyck
President

Board of Directors

A regular meeting of the Board of Directors was held on September 2, 1942. Those present were A. F. Van Dyck, president; C. C. Chambers, I. S. Coggeshall, Alfred N. Goldsmith, O. B. Hanson, F. B. Llewellyn, Haradan Pratt, B. J. Thompson, H. M. Turner, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

George Lewis was appointed official representative of the Institute in England for the duration of the war.

The Winter Convention which would normally be held in New York City in January, 1943, was canceled.

A recommendation of the Executive Committee that a New York Section be established and the necessary procedure to do so were approved.

During the past several months, the Secretary has devoted a substantial part of his time under Institute authorization to a project sponsored by the War Production Board on the preparation of standards covering radio components for the Military Services, which project is under the administrative direction of the American Standards Association. It was agreed by the Board that any required proportion of the Secretary's time would be made available for this work by the Institute during the next several months. This action was taken by the Board because of the vital importance of the project to the prosecution of the War.

Executive Committee

The Executive Committee met on August 18 and those present were A. F. Van Dyck, chairman; Alfred N. Goldsmith, R. A. Heising (guest), F. B. Llewellyn, Haraden Pratt, B. J. Thompson and H. P. Westman, secretary.

At the recommendation of Advertising Manager Copp, the advertising rates for the PROCEEDINGS were revised to provide for a 15 per cent discount to the agency placing the advertisement.

Several letters of petition naming additional candidates for the office of Director were received. An insufficient number of signatures were affixed to those petitions received before August 15 to comply with the Constitutional requirements and necessarily the petition could not be received.

We were informed by the Institution of Radio Engineers (Australia) of the following resolution passed by the Council of that organization:

"That members of the Institute of Radio Engineers (United States of America) in the fighting forces, resident for the time being in Australia, be invited to become Honorary Members of the Institution of Radio Engineers (Australia) and that a suitable letter be forwarded to General MacArthur, the Institute of Radio Engineers (United States of America), and branches of this Institution advising them of the above decision of the Council."

Members of the Institute desiring to partake of these privileges may communicate directly with the Institution of Radio Engineers, Science House, 157 Gloucester Street, Sydney, N. S. W., or with any of the following division secretaries: K. L. Elliott, Brisbane Division, Box 1765 W, General Post Office, Brisbane, Queensland; R. R. Mackay, Melbourne Division, c/o Melbourne Technical College, 134 Latrobe Street, Melbourne, Victoria; and S. C. Austin, Perth Division, Box 335, General Post Office, Perth, Western Australia.

The Institute extends a cordial invita-

Roy Alexander Weagant
1881–1942

Roy Alexander Weagant (M'13-F'15) was born on March 29, 1881, in Morrisburg, Ontario, Canada. When he was but a few years old, his family moved to Dernby Line, Vermont.

He received a Bachelor of Science degree from McGill University in 1905. The next year was spent with the Montreal Light, Heat, and Power Company, and the following year with the Westinghouse Electric and Manufacturing Company at East Pittsburgh, Pennsylvania. In 1907, he entered the employ of the DeLaval Steam Turbine Company of Trenton, New Jersey. From 1908 to 1915 he was with the National Electric Signalling Company. He became Chief Engineer of the Marconi Wireless Telegraph Company from 1915 to 1920 and served as a Consulting Engineer for the next four years to the Radio Corporation of America which was a successor to the Marconi Company. He then served as Chief Engineer and Vice President of the DeForest Radio Company and later devoted his time to a consulting practice.

Mr. Weagant was the second recipient of the Morris Liebmann Memorial Prize which he received in 1920 for his work on methods for reducing the effects of static in the reception of radio signals.

The death of Mr. Weagant, which occurred at Newport, Vermont, on August 23, 1942, marks the passing of another of the early pioneers who contributed substantially to the new art of radio communications.

tion to all members of the Institution of Radio Engineers (Australia) who are located in the United States to attend its meetings and to avail themselves of such facilities as the Institute office can provide.

It was recommended to the Board of Directors that George Lewis, who has been recently ordered to London for the duration by the International Telephone and Telegraph Corporation, be designated an official Institute representative in England.

It was noted that a number of conventions including the Rochester Fall Meeting for 1942 had been canceled and in view of the limited attendance and difficulties in obtaining suitable papers for the summer convention in Cleveland, it was recommended to the Board of Directors that the Winter Convention which would normally be held in New York in January, 1943, be canceled.

An invitation from the Chicago Section that the Institute convene there in June, 1943, was considered. In view of the unsettled conditions, no action was taken on it.

All members of the Executive Committee were present at a meeting held on August 26. The attendance included A. F. Van Dyck, chairman; I. S. Coggeshall, Alfred N. Goldsmith, R. A. Heising (guest), F. B. Llewellyn, Haraden Pratt, B. J. Thompson, and H. P. Westman, secretary.

Dr. Llewellyn, who is that member of the Executive Committee charged with the responsibility for overseeing the activities of our technical committees, reported on a meeting of the chairmen of the various technical committes, and, as a result, the following three actions were taken.

The terms of service of all technical committees is the calendar year at the present time. This results in two major periods of inactivity each year. One of these is during the summer months, and the other is during the first couple of months of the year when committee personnel is appointed. It was agreed that it would be desirable to start the term of service in the early summer and thus avoid the lull which occurs during the winter when the committees should be in full activity. This requires a modification of the Bylaws and Dr. Llewellyn will prepare a proposed revision for submission to the Board of Directors.

It was agreed that brief or partial standards reports may be published in order that they might appear at an earlier date than would be possible if final reports were required. The diminution in activities of committees as a result of the war effort may make it impossible to prepare complete reports in some instances.

With the exception of a few specified committees, the secretarial assistance normally provided to all Institute committees will be restricted to the preparation and mailing of notices of meetings and of duplicating and distributing minutes and reports of the meetings. This action, which applies to all technical committees and most of the administrative committees, was considered necessary in order to release the headquarters staff and the Secretary for other duties.

On September 1, A. F. Van Dyck, Chairman; Alfred N. Goldsmith, F. B. Llewellyn, Haraden Pratt, B. J. Thompson, and H. P. Westman, Secretary, attended a meeting of the Executive Committee.

Malcolm Parker Hanson
1894–1942

Malcolm Parker Hanson (A'21-M'29) was born of American parents in Berlin, Germany, on October 19, 1894.

After three years at the University of Wisconsin he enrolled in the United States Naval Reserve Force in 1917 as a radio electrician. Later, he was commissioned as an Ensign and in June, 1919, he was released from active service.

In 1924, as a civilian engineer, he joined the staff of the Naval Research Laboratory, subsequently being placed in charge of the Aircraft Radio Section. He acted as technical advisor on radio matters to the Wilkins-Detroit, the Navy-MacMillan, and the Byrd Antarctic expeditions, and to the Byrd North Pole and transatlantic flights. He accompanied Admiral Byrd on his 1928–1930 expedition to the Antarctic.

On his return he was placed in charge of the technical work at the Radio Flight Test Laboratory of the Naval Air Station at Anacostia.

In 1938, he became Vice President of the Radio Navigational Instrument Corporation in New York, manufacturers of radio navigational instruments for aircraft.

In 1939, he returned to the Navy as a Lieutenant Commander and was promoted to Commander two years later.

In recognition of his work he was awarded a special gold medal by Congress in 1930; the gold medal of the Veteran Wireless Operators' Association which was awarded by radio from New York to Little America; the special gold medal from the City of New York, and the medal of the Aeronautical Chamber of Commerce.

Commander Hanson's death occurred in the line of duty in Alaska on August 9, 1942, where he was killed in an airplane accident. With his death, the radio engineering profession lost one of its keenest minds and principal authorities on all phases of aircraft radio.

Approval was granted of 94 applications for Associate, 6 for Junior, 29 for Student, and 2 for transfer to Associate grade.

The Secretary reported that under date of August 27, 1942 the PROCEEDINGS was admitted to membership in the Audit Bureau of Circulations.

It was recommended to the Board of Directors that as great a proportion as may be required of the Secretary's time be made available to the American Standards Association for activities on a Government-sponsored project to develop standards on radio components for the Military Services.

It was recommended to the Board of Directors that a New York Section be formed and a procedure to bring this about was approved.

Dr. Schelkunoff (right) receiving from President Van Dyck the check for the Morris Liebmann Memorial Prize for 1942. The presentation was made at the Summer Convention Banquet on June 30 in recognition of Dr. Schelkunoff's contributions to the theory of electromagnetic fields in wave transmission and radiation.

Summer Convention

Our summer convention for 1942 was held in Cleveland, Ohio, on June 29, 30, and July 1. The program was published in the May, 1942, issue of the PROCEEDINGS and the following papers which were not contained in that announcement were among those presented.

"Half-Wave Voltage-Doubling Rectifier Circuit," by W. D. Waidelich and C. H. Gleason of the University of Missouri, Columbia, Missouri.

"A Solution of the Problem of Adjusting Broadcast Directional Arrays With Towers of Unequal Heights," by J. M. Baldwin and G. H. Brown, KDYL, Salt Lake City, Utah, and RCA Manufacturing Company, Camden, N. J., respectively.

The paper on "Stub-Feeder Calculations," by H. A. Brown and W. J. Trijitzinsky of the University of Illinois was not presented through inability of the authors to be present.

A symposium on "What Radio Means in the War Effort" was also included in the program. The following speakers participated.

A. F. Van Dyck, President, Institute of Radio Engineers, Chairman.

Paul Galvin, President, Radio Manufacturers Association.

Neville Miller, President, National Association of Broadcasters.

E. K. Jett, Chief Engineer, Federal Communications Commission.

E. M. Webster, Captain, United States Coast Guard.

Glen Bannerman, President of the Canadian Association of Broadcasters, was originally scheduled to speak but was unable to be present.

The material presented by the Presidents of the Institute, the Radio Manufacturers Association, and the National Association of Broadcasters appears elsewhere in this issue.

Captain Webster discussed the radio problems which face the Coast Guard in its wartime activities.

The work of the Federal Communications Commission was outlined by Mr. Jett. He discussed the problems which the advent of war had brought about in the realm of radio regulation. The problems of making greatest use of the radio spectrum for the war effort were considered and included not only the need for effective point-to-point communications but also the necessity of maintaining public morale through a continuation of broadcasting. The problems of monitoring such a large spectrum as radio now encompasses, to prevent inimical use of radio, is a task of great proportions on account of the geographical size of the United States, the characteristics of radio waves, and the ability to carry on long-distance communications with relatively low powers. The need for full co-operation by all radio organizations and individuals will be required to give the most effective use of this important medium of communication in winning the war.

During the banquet, the Morris Liebmann Memorial Prize was presented to Dr. S. A. Schelkunoff for his contributions to the theory of electromagnetic fields in wave transmission and radiation. About 150 members and guests attended the banquet.

At the luncheon on June 30, Frazier Hunt, news reporter and correspondent, spoke on the war and the immediate probabilities which might be expected from the conditions then existing. His appearance was through the courtesy of the General Electric Company.

The registration at the Convention included 256 men and 36 women.

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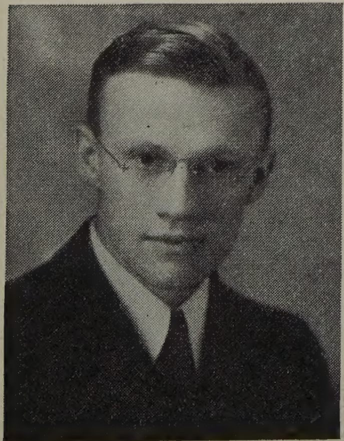
Contributors

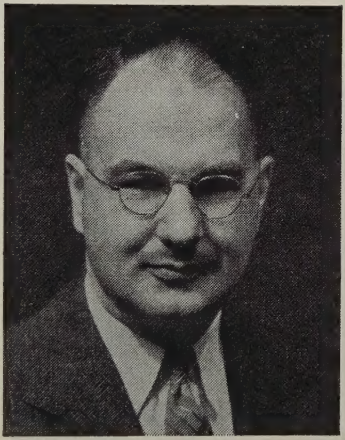
Alda V. Bedford

Alda V. Bedford (A'31) was born in Winters, Texas, on January 6, 1904. He received the B.S. degree in electrical engineering from the University of Texas in 1925. While at the University he spent one summer with the Dallas Power and Light Company, and during the latter part of his school term he was engaged as assistant in the physics department. In 1925 Mr. Bedford joined the General Electric Company, starting in the general engineering department and later transferring to the testing department and research laboratories, working on sound recording by film and disk, audio-frequency amplifiers, loudspeakers, sound printers for film, and television. While in Schenectady he obtained the M.S. degree in electrical engineering from Union College. Since 1929 he has been employed in the laboratories of the RCA Manufacturing Company, first on disk sound recording and then on television. He received a "Modern Pioneer" award from the National Association of Manufacturers in February, 1940, for inventions in television.



G. L. Fredendall

Charles M. Burrill (A'24–M'30) studied electrical engineering at the University of Minnesota, graduating in 1923. He then went with the General Electrical Company. Following three years of general training in the General Electric advanced course in engineering, he joined the radio engineering department, and in 1927 was placed in charge of tuned-radio-frequency receiver development. Since 1930 Mr. Burrill has been with RCA at Camden, N. J., with the exception of a year and a half in 1931–1932 spent with the Rogers-Majestic Corporation of Toronto, Canada, in charge of research. Since returning to Camden he has been engaged in general research, first in sound recording, and more recently in interference and noise suppression and in ultra-short-wave propagation. He is a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and the Franklin Institute.



Charles M. Burrill

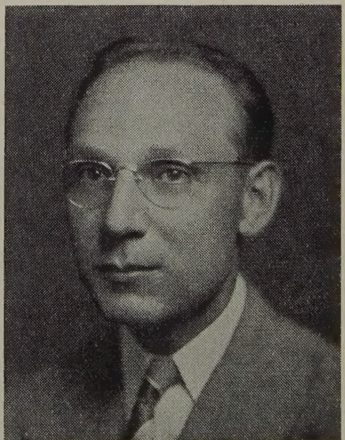
G. L. Fredendall (A'41) received the Ph.D. degree from the University of Wisconsin. From 1931 to 1936 he was at the University teaching electrical engineering, mathematics, and doing research work in mercury-arc phenomena. Since 1936 he has been with the RCA Manufacturing Company engaged in television research.

Irvin E. Fair (A'39) was born at Iola, Kansas, on June 30, 1907. He received the B.S. degree in electrical engineering from Iowa State College in 1929. The same year Mr. Fair entered the radio research department at the Bell Telephone Laboratories and has been engaged principally in the study of piezoelectric crystals and crystal oscillators.

Martial A. Honnell

Martial A. Honnell (A'40) was born on October 23, 1910, at Lyons, France. From 1928 to 1930 he was a shipboard radio operator with the Radiomarine Corporation of America. He received the B.S. degree in electrical engineering in 1934 and the M.S. degree in electrical engineering in 1940 from the Georgia School of Technology. Mr. Honnell was on the engineering staff of WGST from 1930 to 1936, and worked part-time for the Van Nostrand Radio Engineering Service of Atlanta in 1935. He was with the radio division of the Pan American Airways at Miami in 1936–1937. From 1937 to 1941 he was instructor of electrical engineering at the Georgia School of Technology, advancing to assistant professor in 1941, and is now head of the communication division of the electrical engineering department. He is a member of Tau Beta Pi and Eta Kappa Nu.

Ray D. Kell (A'35) received his B.S. degree in electrical engineering from the University of Illinois in 1926. From 1926 to 1930 he was engaged in television research in the radio consulting laboratory



Irvin E. Fair

Ray D. Kell

of the General Electric Company. From 1930 to the present time he has been a member of the research division of the RCA Manufacturing Company, where he has continued his work on various television problems. He received a "Modern Pioneer" award from the National Association of Manufacturers in February, 1940, for inventions in television. Mr. Kell is a member of Sigma Xi.

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Henry N. Kozanowski (A'34), was born on August 15, 1907, at Buffalo, New York. He received the B.S. degree from the University of Buffalo in 1927; the M.A. degree in 1929; and the Ph.D. degree in physics from the University of Michigan in 1930. Dr. Kozanowski held a gradute teaching assistantship at the University of Buffalo during 1927 and 1928. He was a research assistant at the University of Michigan during 1929 and 1930. From 1930 to 1935 he was in the research laboratories, power tube section, of the Westinghouse Electric and Manufacturing Company. Since 1935 he has been in the research laboratories of the RCA Manufacturing Company. He is a member of Sigma Xi, Phi Beta Kappa, and the American Physical Society.

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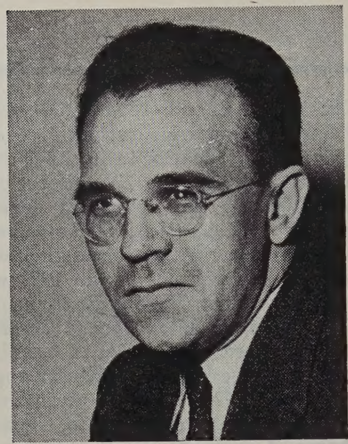
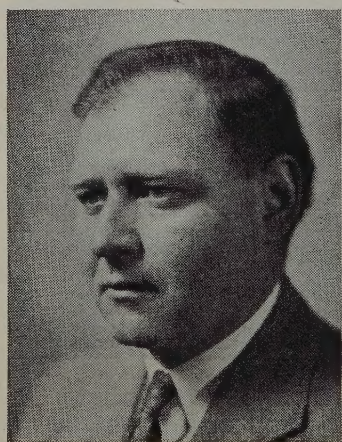


Henry N. Kozanowski

Warren P. Mason (A'36-F'42) was born in Colorado Springs, Colorado, on September 28, 1900. He received the B.S. degree in electrical engineering from the University of Kansas in 1921, the M.A. degree from Columbia University in 1924, and the Ph.D. degree in 1928 from Columbia. He has been a member of the research department of the Bell



Warren P. Mason

W. E. Ruder

Telephone Laboratories since 1921. His work has been mainly with wave propagation networks, both electrical and mechanical, and with piezoelectric crystals. Dr. Mason is now head of the department investigating piezoelectric crystals. He is a member of the Physical Society and a Fellow of the Acoustical Society.

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W. E. Ruder was born in Stockdale, Pennsylvania. He is a graduate of Southwestern State Normal School and of Pennsylvania State College.

Mr. Ruder has been with the General Electric Company since 1907. In 1920 he was made head of the magnetic section of the research laboratory, and has been head of the metallurgical and magnetic section since 1938. He is the holder of many patents on magnetic and metallurgical processes and products, such as calorizing, alnico magnets, resistance alloys, magnetic sheet materials, etc., and is the author of many papers on metallurgical and magnetic subjects. He is a director of the Allegheny Ludlum Steel Corporation and a member of Theta Xi Fraternity, American Iron and Steel Institute, American Society for Metals, American Association for the Advancement of Science, and the American Institute of Mining and Metallurgical Engineers.